Proceedings of the I·R·E

NOVEMBER 1943

VOLUME 31 NUMBER 11

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F-M Transmitter Receiver

Power-Tube Performance

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Radio Production

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MEMBER AUDIT BUREAU OF CIRCULATIONS

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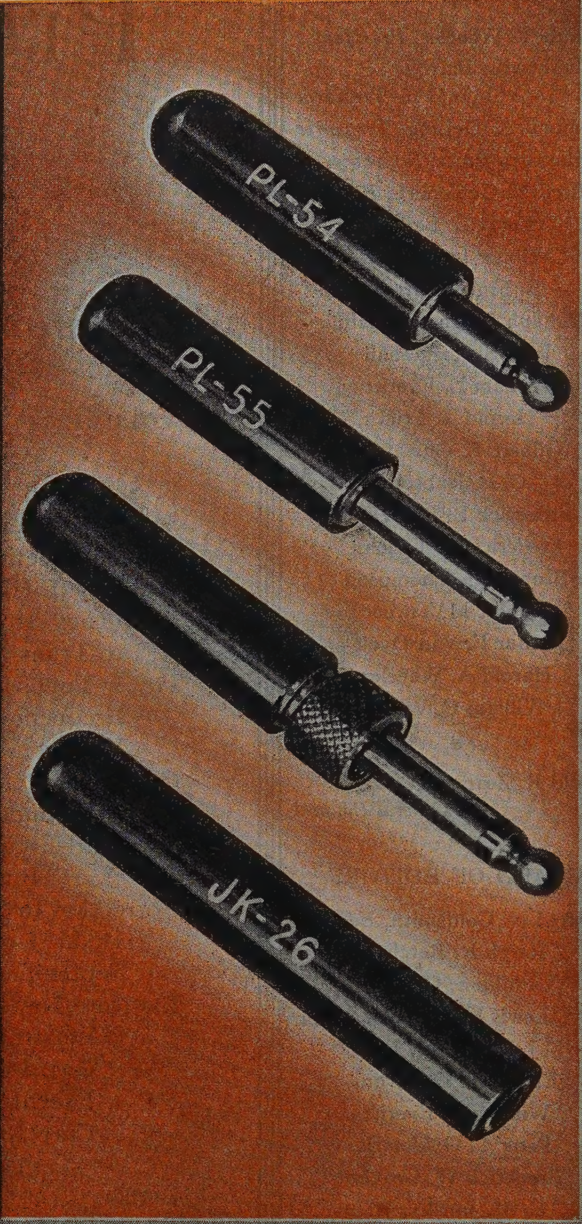

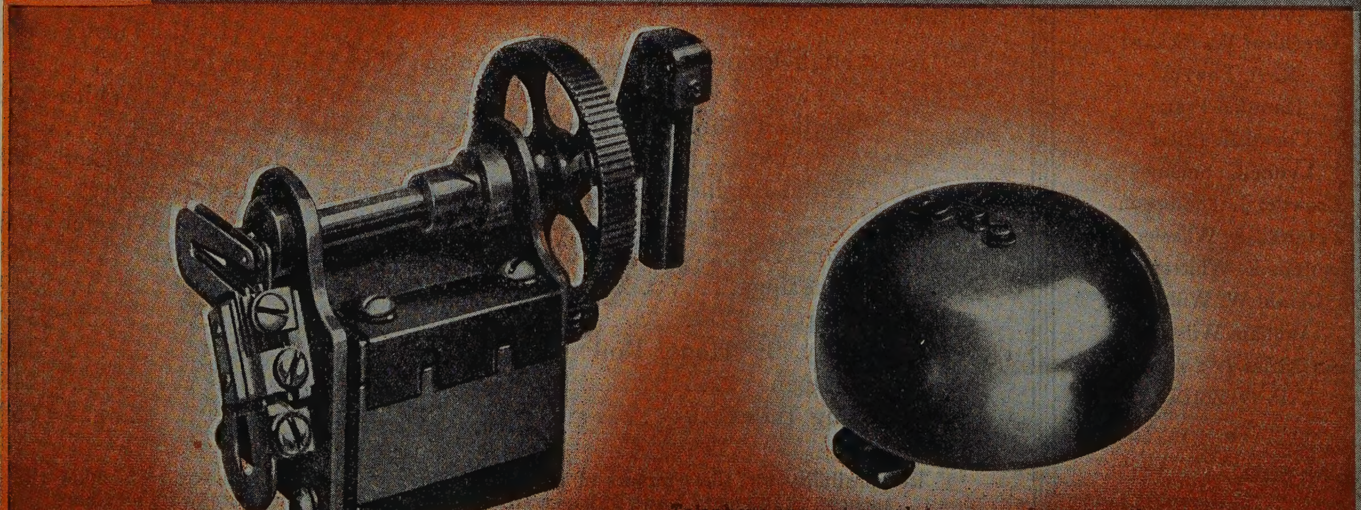

Telephone generator and ringer are shown less than actual size.

F. S. Barton

F. S. Barton is the eldest son of the late Professor E. H. Barton, D.Sc., F.R.S., of Nottingham, who was engaged on radio research work under Hertz at Bonn in 1892. Mr. Barton, who was born on May 13, 1895, started life in an atmosphere of radio research. His subsequent education at St. John's College, Cambridge, of which he was a Foundation Scholar, was interrupted by World War I. In the earlier part of the war, he was engaged in munitions research, involving a mission to Russia; in August 1917 he was commissioned a Sub-Lieutenant R.N.V.R. for radio experimental duties with the Royal Naval Air Service. Transferring to the R.A.F. on its formation, he was finally demobilized as a Captain, R.A.F., in May, 1919.

Returning to Cambridge to resume his physics course under Lord Rutherford, Mr. Barton received the degrees of B.Sc. (War) London and M.A. Cantab, and then rejoined the Royal Air Force in a civilian capacity as a member of the Instrument Design Establishment, later transferring to the Royal Aircraft Establishment, South Farnborough.

While at the Royal Aircraft Establishment, Mr. Barton was concerned with the research side of communication equipment for the Royal Air Force and was especially interested in the development of radio-controlled target aircraft and radio aids to blind landing. In this latter connection, he visited a number of European countries and the United States in 1934–1935, making a survey of the problem for the Civil Aviation Department of the Air Ministry.

Mr. Barton was Air Ministry representative on various committees of the Radio Research Board, Department of Scientific and Industrial Research, and of the Wireless Committee of the Institution of Electrical Engineers and on various military committees and boards. During this time he was elected a Member of the I. E. E. and a Fellow of the Institute of Physics. In 1935 he was elected a Fellow of the Institute of Radio Engineers and gave an address on "Enemy War Radio" to a New York meeting in October 1942.

In June, 1941, Mr. Barton was transferred to the British Air Commission, Washington, as Chief of the Radio Division, and was elected Vice President of the Institute of Radio Engineers for the year 1943.

Radio-and-electronic engineers are deeply concerned with the trends of the industry to which they devote their efforts. Expressions of opinion of commercial leaders of the industry, published in the PROCEEDINGS in the form in which they are received, will serve to clarify engineering ideas and to direct technical thought and effort. To these ends, there is here presented an analytic "guest editorial" by the President of the Belmont Radio Corporation.

The Editor

Some Comments on Postwar Electronics

P. S. Billings

The fabulous wartime expansion of the electronic industry has created a veritable host of prophets, crystal-ball gazers, and calamity howlers. Some are sincere, some patently dishonest, but all contribute to the almost unbelievably fantastic public misconception of the part electronics will play in the immediate postwar period. Ambiguous and in some cases deliberately misleading advertising, picturing and depicting postwar radio, has only added to the confusion. Overzealous advertising is building the public up to an anticlimax, the reverberations of which may well shake the very foundations of our embryonic industry. The public will expect promises made now to be kept in the immediate postwar period and "gilding the lily" will only lead to a retarded market which may well snuff out many potential postwar developments. The fantastic advertising efforts to capture the public imagination seem to have been engendered by the fear that industry has overexpanded and that only through the capture of a lion's share of the postwar market can the individual companies hope to survive in the pristine glory to which they have become accustomed.

Prior to the war, the electronic industry was already on its way to a peacetime expansion previously unequalled in any but the automotive industry. The war itself served to accelerate this expansion principally through the diversion of engineering talent ordinarily devoted exclusively to the broadcast receiver and associated branches of the industry into lines of endeavor not previously emphasized or exploited, as either commercial or industrial. In addition the engineering staffs of our foremost seats of learning turned their efforts from teaching and academic research to design and practical development. It is to be emphasized that the war has served to train our sights upon the reduction to practice of much fundamental research hitherto allowed to be dormant and neglected. This effort to practicalize and find applications for hitherto obscure and little-known phenomena has led to a precocious growth which otherwise would have been delayed for many years. This growth is most fortunate since otherwise drastic reduction of facilities and personnel would have been required in the postwar conversion.

In looking forward to the postwar electronic picture we may briefly summarize the effect of the war on the various branches of electronics as follows:

Radio broadcasting will be in practically the same state as prior to the war. Frequency modulation has not advanced technically even though it found widespread military application. While unquestionably frequency modulation will find an almost immediate public acceptance, this acceptance was

well on the way to establishment before Pearl Harbor. The opening up of the ultra-short-wave and microwave spectra to other services may result in an indirect benefit through possible expansion of the band now allocated to this service.

While there has been very little direct television development since the start of the war, it has received many indirect benefits through the close similarity of its technique with that of certain allied fields. If, however, television is to receive the benefits of these developments the release of television to the public must be delayed for several years in order to permit the redesign of the circuits for much higher frequency operation and to make other changes which appear advisable in view of advances in pulse technique, modulation methods, and cathode-ray equipment. This delay might also enable possible simultaneous release of color television.

Facsimile has received some direct benefit from the wartime development. It resembles television, however, in that much postwar development is required before it can be released to the public.

In the case of both television and facsimile a major merchandising and financial problem must be solved before an appreciable growth can be expected.

Navigational aids and equipment have received a tremendous impetus. The application of radio to the navigation of air and surface craft will become universal. New positioning devices will replace existing direction-finding equipment. Airway equipment will probably be moved into the ultra-short-wave spectrum. The development of the helicopter and its probable widespread public distribution makes this branch one which will require extensive engineering and production facilities.

The strictly communication field has been greatly expanded through the development of ultra-short and microwave technique for both mobile and specialized point to point communication. This technique will unquestionably find large application in the control of common carriers and other mobile units hitherto impractical because of the limited usable frequency spectrum.

Industrial electronics have just begun to scratch the surface. With the end of the war and with more capable engineering talent available for this specialized branch, it seems reasonable to assume this specialization will require a substantial portion of the facilities of the electronics field. To enumerate or attempt to classify all the possible electronic applications would require far more space than is here available. It is sufficient to say that a legion of applications are either now already developed or in the process of development. A roster of electronic applications becomes so formidable and impressive that one must conclude that here is one industry which has not been overexpanded by the war effort.

These new applications of electronics are not for the most part in shape for release to the public. During the past two years, however, the manufacture of home radio has been cut off. Manufacturers can use the consequent pent-up demand for new equipment as a stopgap to hold their expanded production organizations intact while all possible pressure is placed on the completion and finishing off of the development of the newer members of the electronic family.

With sane and sensible planning for the conversion to peacetime status, it will not be necessary to curtail either personnel or existing facilities. Hysterical attempts to capture the public fancy on the basis that there will not be enough business to go around are not only unnecessary but unwise. The saner policy consists in a planned and stabilized expansion into the hitherto unexploited fields of our industry. In this expansion you engineers must "carry the ball." The future of our industry is in your very capable hands. Looking postwarward we have every faith in the brightness of your future and of the part that our precocious infant will play in the world to come.

The Institute of Radio Engineers

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ATLANTA	CHICAGO	CLEVELAND	DETROIT
November 19	November 19	November 25	November 16

LOS ANGELES	NEW YORK	PHILADELPHIA	PITTSBURGH	WASHINGTON
November 16	December 1	December 2	December 6	December 6

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Stability in High-Frequency Oscillators*

R. A. HEISING†, FELLOW, I.R.E.

Summary—**This paper discusses frequency stability with change in plate voltage of high-frequency oscillators of around 100 megacycles and shows both theoretically and experimentally that the highest stability found by many is only the result of fortuitous circuit adjustment that may readily lead to the desired result in this frequency range. It is shown that the factor next in importance in producing frequency stability is a low ratio of inductance to capacitance in the frequency-determining circuit. It is also shown that a high Q contributes little directly to stability. A high Q is necessary with low L/C ratios to get oscillations but an improvement in Q alone may give poorer stability. To get the fullest measure of stability with low L/C and high Q calls for slight adjustments in the circuit and possibly the provision of loose coupling to the frequency-determining circuit.**

Introduction

FREQUENCY stability with plate-voltage variation in electric oscillators has been the subject of many investigations in the past twenty years, and has given rise to a number of papers. The subject in most of its aspects has been well covered and such phases will, therefore, not be treated here. However, there has developed in the last few years a field wherein the electric oscillator has held its own in competition with crystal control and for which there is still need of further study. This field is the ultra-high-frequency range from 30 to 300 megacycles for which crystals have not been cheaply and easily applicable.

In the published papers on stabilized ultra-high-frequency oscillators, the investigators have described performance which equalled or surpassed what could be obtained with crystal oscillators. The experimental experiences and theoretical treatments published, however, have not been adequate for properly designing such oscillators or proportioning the parts. It was found that when the oscillator had been designed and constructed, improved stability could always be secured by further adjustment of the constituent elements. It was further observed that most, if not all, of these published stabilized oscillators were brought to their points of maximum stability by such adjustments, and no criterion for such adjustment has been given. It was felt, therefore, that the underlying principles for these stabilized oscillators had not yet been fully identified.

In discussing the subject it appears desirable in the interest of clearness first to state three observations. The first observation for which supporting evidence has been found involves the general mechanism whereby maximum stability is attained. As mentioned previously, it is usually found that the best stability of ultra-high-frequency oscillators is finally obtained by adjustment of circuit elements. Even the value $df/dE=0$ can be secured by such adjustment. This leads to:

Observation I—That the maximum stability in ultra-high-frequency oscillators is produced by a fortuitous adjustment easily brought about due to the respective interelectrode impedance relations of the tube obtaining at ultra-high frequencies.

Experimental and theoretical reasons will be supplied to support a second observation, viz.,

Observation II—That the circuital design next in importance in producing high stability is a low ratio of L/C in the frequency-determining circuit.

Because of variance from statements occurring in the literature, a third observation is given herewith:

Observation III—That a high Q in the frequency-determining circuit contributes little or nothing *alone* to stability and may actually reduce it, but when combined with proper circuital changes will give an improvement.

It should be understood that the material in this paper applies only to oscillator circuits of the simple well-known regenerative types where the frequency-determining circuits are connected between tube electrodes and operate at frequencies in the lower end of the ultra-high-frequency range.

Mathematical studies of oscillators show several methods of improving stability.[1] It is not always possible to pick beforehand the method of stabilizing which one wishes and then proceed to proportion the elements to produce the desired result. Too often the large number of independent and dependent parameters are so interwoven that the desired values or rates of variation cannot be predetermined. It is believed that, in so far as ultra-high-frequency oscillators are concerned, an elementary analysis combined with experimental evidence will indicate the most important principles underlying their stable performances.

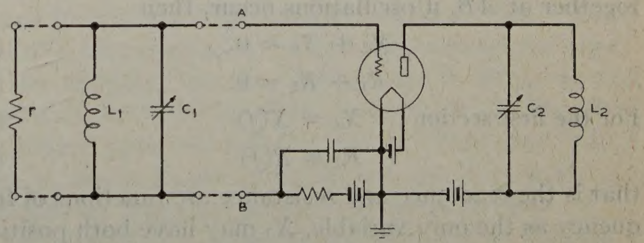

Fig. 1—Circuit used in the test. L_1C_1 is the circuit whose constants were varied, and r is the resistance used to change its Q. Grid leak resistance 5000 ohms. $E_g=-30$ volts. Vacuum tube 316A.

In Fig. 1 there is shown a common type of oscillator circuit. The circuit, ignoring an extra resistance r, is divided into two sections. The first section, that to the left of AB, contains only the inductance and capacitance of the input-grid tuned circuit which has effective

* Decimal classification: R355.9. Original manuscript received by the Institute, March 20, 1943; revised manuscript received, May 28, 1943.

† Bell Telephone Laboratories, Inc., New York, N. Y.

[1] F. B. Llewellyn, "Constant frequency oscillators," Proc. I.R.E., vol. 19, pp. 2063–2095; December, 1931.

resistance R_1 and reactance X_1 at the oscillating frequency as measured at the two terminals. This is the circuit which in this treatment embodies those elements that exert the major frequency-determining effort. In this circuit are placed the high-Q elements in determining their relation to frequency stability. The remainder of the oscillator circuit, that to the right of AB, including the tube, constitutes the second section and measures resistance R_2 and reactance X_2 at its terminals at the oscillating frequency with some certain amplitude of applied voltage. As is well known from the published literature on oscillators, any oscillator circuit may be broken into at any point, and if the resistance and reactance are measured at the natural oscillating frequency with a current amplitude equal to that which obtains as the circuit oscillates, it will be found that both the resistance and reactance are zero. Therefore,

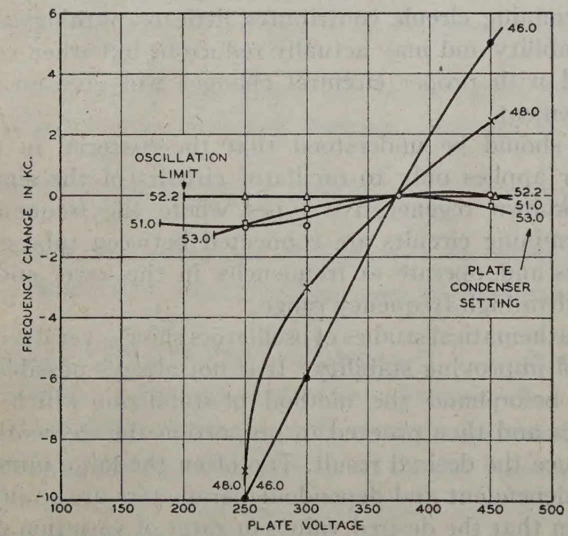

Fig. 2—Curves of frequency as a function of plate voltage with various settings of the plate tuning condenser. Both positive and negative slopes are present. The slope of such curves at the operating potential is termed the "instability." Zero instability is therefore zero slope or maximum stability.

when the circuit of Fig. 1 has its two parts connected together at AB, if oscillations occur, then

$$X_1 + X_2 = 0 \tag{1}$$

$$R_1 + R_2 = 0. \tag{2}$$

For the first section
$$X_1 = X(f) \tag{3}$$

$$R_1 = R(f) \tag{4}$$

that is the reactance and resistance are functions of frequency as the only variable. X_1 may have both positive and negative values, while R_1 will be positive only. For the second section,

$$X_2 = X(E, A, g_m, f) \tag{5}$$

$$R_2 = R(E, A, g_m, f) \tag{6}$$

that is the reactance and resistance vary due to variations in E (plate voltage), A (amplitude of voltage on grid), g_m (mutual conductance of tube), and f (frequency). X_2 may have positive and negative values in different parts of the frequency range but will be negative at the oscillating frequency. R_2 may also have positive and negative values in different parts of the frequency range and will also be negative at the oscillating frequency. Since (1) must hold for all variations in the circuit, one can write

$$dX_1 + dX_2 = 0 \tag{7}$$

and expanding to take account of the variable parameters

$$\frac{\partial X_1}{\partial f} df + \frac{\partial X_2}{\partial f} df + \frac{\partial X_2}{\partial A} dA + \frac{\partial X_2}{\partial E} dE + \frac{\partial X_2}{\partial g_m} dg_m = 0. \tag{8}$$

Rearranging

$$\frac{df}{dE} = -\frac{\dfrac{\partial X_2}{\partial E} + \dfrac{\partial X_2}{\partial A}\dfrac{dA}{dE} + \dfrac{\partial X_2}{\partial g_m}\dfrac{dg_m}{dE}}{\dfrac{\partial X_1}{\partial f} + \dfrac{\partial X_2}{\partial f}} \tag{9}$$

which expresses the variation of frequency with plate voltage in terms of the other parameter variations.

It now becomes necessary to show in what way the stability equation (9) agrees with experimental facts. Fig. 2 shows five curves for an ultra-high-frequency oscillator having the circuit of Fig. 1 with five different settings of the plate tuning condenser. The input tuned circuit was constructed to have a high Q. It is to be observed that positive as well as negative and zero slopes are obtained.

In reconciling these experimental curves with stability equation (9) one should note first the experimental fact that the rate of variation of frequency with plate voltage can be made zero. That is

$$df/dE = 0. \tag{10}$$

To make (9) equal to zero requires either that the numerator be zero or that the denominator be infinite. The denominator can become infinite only if one of the terms becomes infinite. Inasmuch as the opinion is widely held that a high-Q circuit is the basis of stabilization this phase of it may be investigated first. By making input circuit reactance X_1 with a high Q, a high value of $\partial X_1/\partial f$ may be obtained at some values of X_1 and f. However, an infinite value can be obtained only if the resistance is actually zero, which is not possible. Making $\partial X_1/\partial f$ large by improving the Q, therefore, does not account for the stabilization.

In addition, positive slopes of the curve are to be found experimentally as shown in Fig. 2, which would require in (9) a negative value for $\partial X_1/\partial f$. A negative value has been pointed out in the literature[2] as an unstable condition. A circuit of the type indicated in Fig. 1 will allow of oscillations only where X_1 is positive and $\partial X_1/\partial f$ is positive. For these reasons, therefore, (9) cannot be made zero or positive by any practical values assignable to the denominator terms involving the high-Q elements. This proves that the stabilities observed cannot be explained upon the basis of a large $\partial X_1/\partial f$, no matter what its Q.

[2] R. A. Heising, "Audion oscillator," *Jour. A.I.E.E.*, vol. 39, April and May, 1920.

To make df/dE zero in a practical circuit then requires that the numerator of (9) be zero. This is brought about by adjusting the various elements in the second part of the circuit, that is X_2, until a balance is achieved between the positive and negative terms according to principles laid down by Llewellyn.[1] By adjusting the grid leak, or grid-stopping condenser, or variable condenser C_2, etc., X_2 will be altered, and its variation with parameters E, A, f, and g_m can be altered so that no change of frequency is observed with reasonable changes in plate voltage E. This appears to be the type of stabilization obtaining in many high-Q ultra-high-frequency oscillators. Credence is lent to this by statements that have been made by writers when delivering their papers that after inserting the high-Q elements in the circuit, improved stability was secured by readjusting other parts of the circuit. This phenomenon which has been observed many times suggested that in the ultra-high-frequency range, the capacitance reactances between tube elements are so low that together with plate and grid resistances and load impedance they readily produce a situation which will make the numerator of (9) zero. It has been found experimentally that it is relatively easy to stabilize most ultra-high-frequency oscillators, while it is not so easy to stabilize those for longer waves.

Now it is not to be denied that changes in the construction of input reactance X_1 to give a large $\partial X_1/\partial f$ may contribute to stabilization. A change may contribute both directly and indirectly. The direct effect will be considered first.

The term $\partial X_1/\partial f$ contributes directly to stability if it is such as to make the denominator larger. This is the usual conception of its operation. Undoubtedly for a fixed numerator, enlarging the denominator will reduce the slope of a frequency-voltage curve, or will increase the radius of curvature of a curve. The methods of increasing $\partial X_1/\partial f$ must be looked into. Consider first the effect of increasing the Q only. In Fig. 3, curve I represents the reactance curve for a given circuit having constants shown thereunder. Assume that operation occurs at point P_1 on this curve. The slope of the curve at P_1 is the magnitude of $\partial X_1/\partial f$. Assume now that the Q of the circuit is doubled. The reactance curve is now that of curve II. Operation will now occur at P_2 at a slightly different frequency and very *slightly* changed reactance to conform to (1) and (2). The slope $\partial X_1/\partial f$ at point P_2 is, however, only very slightly greater than at P_1. If now the Q should be doubled once more giving reactance curve III, the operating point will be at P_3 and will be so close to P_2 as to be difficult to show on the drawing. The slope $\partial X_1/\partial f$ at that point will change even less. Increasing Q indefinitely thereafter will make little change in position of P or value of $\partial X_1/\partial f$. Change in Q *only* operating through increasing $\partial X_1/\partial f$ cannot give improved stability of any measurable amount, no matter if Q becomes infinite.

However, the first thing that comes to mind after increasing the Q from that of curve I to curve II is to observe that there is a place on the new curve such as point P_4 where the slope actually will be much greater than at P_1 or P_2. Will not a shift to this point give greater stability due to the much larger $\partial X_1/\partial f$? This brings in the indirect effect mentioned. In shifting to point P_4 the magnitude of $\partial X_1/\partial f$ is actually increased markedly. However, such a shift has increased X_1 itself also. The increase in X_1 makes X_2 undergo a decrease since X_1+X_2 must equal zero. X_2 in assuming a new value may have an effect on stability many times as great as the change in the denominator by the new $\partial X_1/\partial f$. Changing X_2 changes the three terms $\partial X_2/\partial E$, $\partial X_2/\partial A$, and $\partial X_2/\partial g_m$ in the numerator. Also, increasing X_1 changes the amplitude of oscillations A upon the grid. It makes the term $\partial A/\partial E$ change. Also, the new amplitude A changes the grid current and grid bias resulting

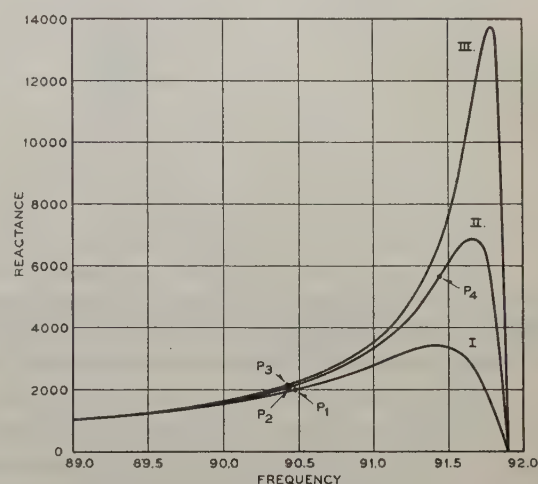

Fig. 3—Reactance curves for an input circuit L_1C_1 in which the Q of the elements only is varied. At the operating points P_1, P_2, and P_3 the slopes $\partial X_1/\partial f$ are practically the same. $L_1 = 120\times 10^{-9}$ henry. $C_1 = 25\times 10^{-12}$ farad. $Q_1=100$. $Q_2=200$. $Q_3=400$.

in a change in g_m and, therefore, of term $\partial g/\partial E$. All in all, the changes in these terms will usually cause a change in the numerator sufficient to outweigh any increase in the term $\partial X_1/\partial f$ in the denominator. If an actual improvement in stability is secured, one is likely to think that improving the Q had produced the result. If the stability becomes worse, one is likely to feel something went wrong while changing the Q. The effect of Q alone upon the stability can be determined only if care is taken to see that X_1 and the numerator are changed by negligible amounts at the same time.

Such stability measurements were made. The input tuned circuit *alone* was altered so as to provide different Q's to the coil with the oscillator operating at the same frequency and with the same X_1. The same identical frequency was essential to insure that no change in X_2 occurred from a change in frequency. In changing the Q, the resistance element used changed the inductance or capacitance slightly so that a slight change in capacitance C was made to restore the frequency. This slight change in L and C and L/C is negligible. However, the main effect of change in Q alone gave anomalous results.

To clarify the relations it was deemed desirable to secure a family of curves in which not only is Q varied but there is varied that element in the circuit which is most likely to be subsequently used in the circuit adjustment. That element is the plate-coupling reactance which in the type of circuit shown resolves into plate-tuning condenser C_2. This element is likely to be adjusted for maximum power, for maximum grid current, for a desired space current, for maximum stability, or for other reasons. To get the family of curves, the Q of the inductance

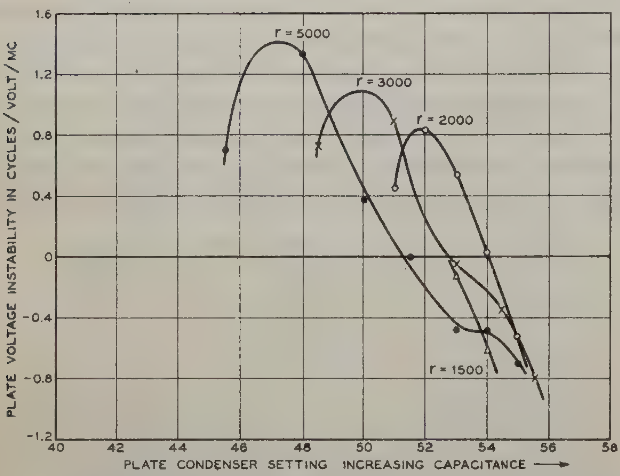

Fig. 4—Instability curves when the Q of the tank circuit is varied by different values of shunting resistance r showing that maximum stability (zero instability) can be secured with several different Q's, and poorer stability may be secured with larger Q's with certain circuit adjustments. Q increases with increase in r.

in the tuned-grid circuit was given different values by connecting across it a high radio-frequency resistance (r in Fig. 1). For each chosen Q, a series of curves of frequency versus plate voltage was plotted for a series of chosen values of C_2 as in Fig. 2. For each chosen Q, C_2 was varied over its entire range for which oscillations would occur. Then from these curves values of $\Delta f/\Delta E$ were measured. It was then possible to plot these values of instability as a function of the adjustment of C_2 for each chosen Q. Such curves plotted in Fig. 4 show, therefore, what can happen to the stability as Q alone is changed for different adjustments in the plate circuit.

The curves show first that maximum stability (intercept on the x axis) is attainable with almost all values of Q for which the oscillator would operate. The curves show second that over certain ranges of C_2 values, an improvement in Q gives greater stability, while over other ranges of C_2 values, improvement in Q gives lesser stability. The curves show third that for high-Q circuits (large r), operation will occur for a wider range of C_2 adjustment and that the wider range brings within its limits greater instability.

The stability, being affected by plate-circuit adjustment, is going to depend upon the aim used in adjustment. If one wishes maximum power in the plate tuned circuit, there is no reason to expect maximum power to occur with the adjustment which gives maximum frequency stability, or if maximum power is desired in the grid circuit, a still different adjustment is likely to occur. It is possible, therefore, if one wants power as well as stability that using a high-Q circuit can lead to a poorer stability after adjustment than if a lower Q were used.

The curves of Fig. 4 are made with such adjustments in the oscillator circuit as to provide maximum stability with some value of C_2. If some element such as grid leak is changed, this set of curves may be raised or lowered as the case may be to the extent of being entirely above or below the X axis. A change in Q only, for a fixed C_2, can then produce opposite changes in stability in the two cases. Similar sets of curves with the grid resistance as the independent variable instead of C_2 can be plotted but they do no more than confirm the evidence that stability increase or decrease with improving Q is purely a matter of the adjustment obtaining in the circuit.

It is, however, possible to operate upon the input tuned circuit and produce desirable effects on stability curves. Such operation is aimed at increasing $\partial X_1/\partial f$ without changing X_1 and with a minimum of disturbance to the numerator of (9). Fig. 5 gives reactance curves for several different sets of inductance and capacitance. Curve I represents the reactance of an arbitrary example of $L=120\times10^{-9}$ henry and $C=25$ micromicrofarads and a Q of 400. The effective value of X_1 at the operating frequency is the ordinate represented by point P. It is now desired to change the circuit so that operation will still occur at point P, but the slope of the curve be increased. By making the ratio of L/C

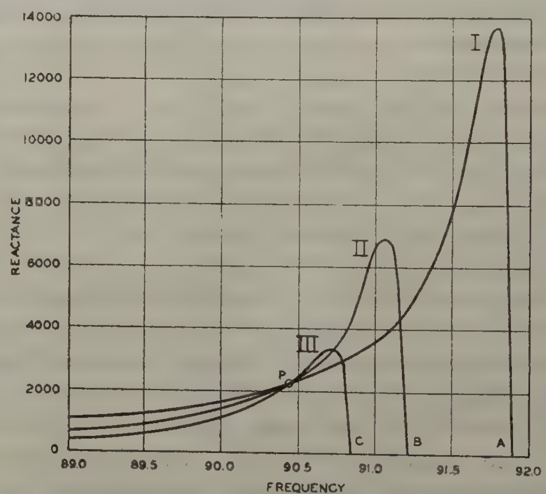

Fig. 5—Curves showing different slopes $\partial X_1/\partial f$ at an operating point P secured by changing the L/C ratio, Q being constant. Curve I, $L_1=120\times10^{-9}$ henry. $C_1=25\times10^{-12}$ farad. $Q=400$. Curves II and III have L/C ratios one fourth and one sixteenth, respectively, that of curve I with a slight shift in the resonant point to give the same frequency in the oscillator circuit.

about one quarter the previous value and shifting the resonant frequency from A to B with Q remaining constant, curve II is obtained passing through P. The slope of this curve at P is greater than for curve I thereby giving the desired increase in $\partial X_1/\partial f$. By reducing the L/C to a quarter the second value, and shifting the resonant frequency to C, still further improvement in $\partial X_1/\partial f$ is obtained. Thus, the denominator of (9) can be increased by reducing the ratio L/C to whatever value is practical to assist in improving the stability.

Further confirmation is secured by constructing an oscillator with a series of input circuits having a wide range in ratio of L/C made with lumped inductance and capacitance, with about the same Q's. Such curves are given in Fig. 6. The circuit with lowest L/C values shows greater stability over the entire operating range. It possesses an inability to be adjusted for as poor stability as for larger L/C values. In these tests, the operating adjustments were so made for the different L/C values that frequency was the same and X_2 was the same at a

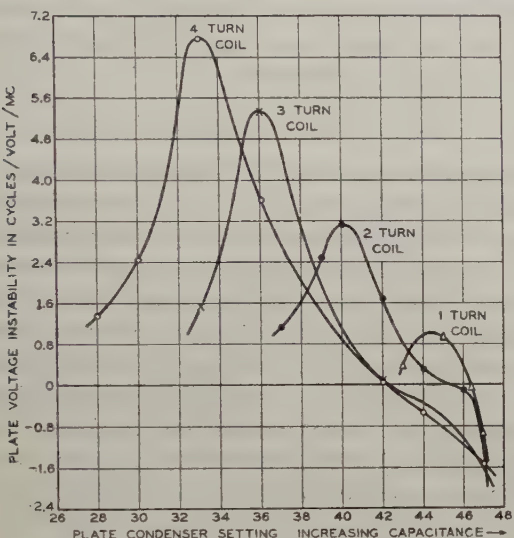

Fig. 6—Instability curves when the L/C ratio is changed, the Q remaining approximately constant. The one with the smallest inductance and lowest L/C ratio will inherently be more stable when adjusted so as to oscillate than the others. All may be adjusted to have the same maximum stability (zero instability) at certain adjusting points.

nominal operating point, thus insuring that the magnitude of X_1 was the same for all L/C values.

An inspection of the published literature shows that the stabilized oscillators described have been constructed with lower L/C ratios than would be the case had lumped inductance and capacitance of previous conventional forms been used. The investigators, in order to get high Q's chose a concentric line or other form of inductance that had low inductance. These investigators unwittingly constructed their oscillators with advantageous L/C ratios following reasoning based upon a false premise. It is probable that they were sometimes puzzled as to why further circuit adjustment gave more stable operation, and also puzzled as to why improved Q circuits sometimes gave less stable operation. If the true basis for the stabilization were known, undoubtedly different designs would have been used in many cases.

Since higher Q's do produce larger $\partial X_1/\partial f$ values on certain parts of the reactance curves, and since $\partial X_1/\partial f$, if large, should contribute to improved stability or give flatter frequency-voltage curves, the question arises as to the proper method of utilizing improved Q. The answer is that looser coupling between the high-Q circuit and the rest of the oscillator must be employed simultaneously. There are a number of factors involved leading to this conclusion. One is the grid-filament effective resistance of the tube. At ultra-high frequencies this becomes low in tubes and when connected across reactance X_1 it can spoil the effective high Q. Another point of view is that this resistance may be effectively introduced into the tuned circuit by the coupling to the tube and if the coupling is too great such introduced resistance may be much greater than the actual resistance. A second reason is that the larger $\partial X_1/\partial f$ occurs only with an increased X_1 (see point P_4 in Fig. 3) and such increased X_1 calls for decreased reactance in the tube's plate circuit. Such necessary plate reactance may thereby be reduced to such a point as not to be able properly to couple the tube to the circuit for power. Also, a large X_1 may require a negative plate reactance at some frequencies which then produces a nonoperable circuit. Furthermore a large X_1 will generate an excessive voltage on the tube grid, wasting power, overheating the tube, and reducing still further the effective resistance. For these various reasons, the magnitude of X_1 must be kept within practical limits and, with high-Q circuits to take advantage of possibilities in securing large $\partial X_1/\partial f$, it is necessary to use loose coupling.

The looser coupling may be obtained by tapping down on the inductance, or placing a small capacitance in series. For example, in the case of greater slope at point P_4 on curve II, Fig. 3, a series-capacitance reactance in the lead to the grid can reduce the total input circuit reactance to the magnitude of P_2 and yet have greater $\partial X_1/\partial f$. In the case of the input circuit used in Fig. 4, of which a cross-section diagram is given in Fig. 7, this "tank" circuit had a Q of 1700. With this Q (and no shunting resistance to reduce the Q) certain adjustments of the plate-tuned circuit would call for high values of

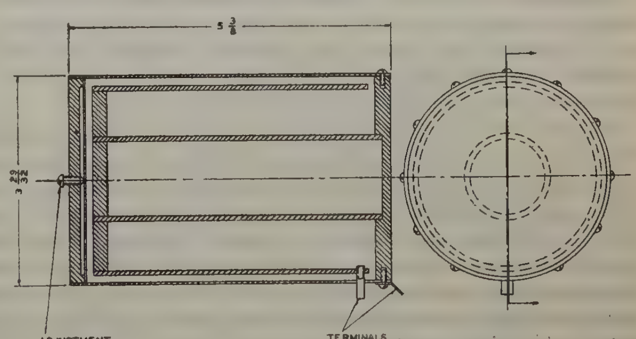

Fig. 7—Drawing showing construction of the tank circuit used in most of the experiments. Inductance approximately 120×10^{-9} henry and capacitance approximately 25×10^{-12} farad.

X_1 that would overload the tube grid. With a series condenser of 9 micromicrofarads, the effective reactance could be brought down to safe values, and at the same time stability curves secured, due to combined high $\partial X_1/\partial f$ and adjustments that were superior to what could be secured without this reduced coupling.

In Fig. 8 are shown several curves in which the tank circuit was given higher values of Q than represented in Fig. 4 by means of higher shunting resistances and where a 9-micromicrofarad coupling condenser was

inserted in the lead to the grid. The abscissas in both cases are plate-tuning condenser settings. The shapes of the curves in the two figures follow the same pattern, the ranges of plate-circuit capacitance over which oscillation will occur are comparable, most of the curves cross the zero axis at some point indicating maximum stability, but the curves of Fig. 8 show less than half

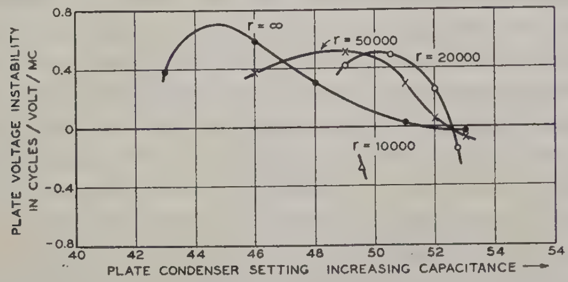

Fig. 8—Curves showing that narrower ranges of instability are secured if a series-capacitance reactance is used with the high-Q tank circuit. 9 micromicrofarads used in these tests. Values of r are shunting values to reduce the effective Q.

the range of instability. *It is the reduced instability range within the entire adjustment range that is contributed by the combination of higher Q and looser coupling.*

The experiments were made with only the one type of oscillator circuit. The results however are applicable to all the more common oscillator circuits.

It is to be observed that with the wide range in Q's and L/C ratios covered, maximum stability could always be secured by circuit adjustment. This checks the observations mentioned by many experimenters. At the ultra-high frequencies, the reactances of the interelectrode capacitances, and of the necessary attached inductances are such that with the tube constants stable circuital conditions are readily obtainable. In lower frequency ranges, in which investigations were carried out and concerning which exhaustive mathematical analyses have been published, such fortuitous circuit conditions did not obtain. It was often found necessary to insert lumped reactances, or unusual-sized circuit elements to secure stabilization. It is hoped that by clarifying in the literature the parts that are played by the elements entering into stability unnecessary or false steps in its attainment may be eliminated.

Conclusions

The experiments show that maximum stability can be secured with a wide range in Q's, the magnitude of the inductance remaining constant, and that it is secured by proper adjustment of the circuit elements. The curves show that with certain circuit adjustments, improving the Q only will give poorer stability. With reduced L/C ratios and constant Q a reduced instability range results, although oscillations will not be secured over as wide a range in misadjustment of a circuit element. With improved Q and looser coupling, a reduction of instability range is also secured.

Frequency-Modulation Transmitter and Receiver for Studio-to-Transmitter Relay System*

W. F. GOETTER†, ASSOCIATE, I.R.E.

Summary—A complete studio-to-transmitter system for high-fidelity program relaying between the studio and the main transmitter is described. The entire equipment was designed considering simplicity and reliability to be of prime importance. Several installations have been in successful operation for over one year.

The 25-watt transmitter incorporates several novel features which account for the excellent performance obtained. Newly designed tubes especially suited for ultra-high-frequency operation are used.

A crystal-controlled, double-conversion superheterodyne receiver, employing such features as cascade limiting, carrier-off-noise suppression, and vertical-chassis construction is also described. Harmonics from the same crystal oscillator are used in performing both conversions, resulting in an extremely stable unit. Both transmitter and receiver may be remotely operated when proper compliance is made with Federal Communications Commission regulations.

A high-gain studio-to-transmitter antenna, which meets all Federal Communications Commission requirements, is totally enclosed against the weather to avoid ice-melting problems.

A frequency-modulation station monitor indicates center frequency continuously, as well as per cent modulation and carrier level. Aural monitoring is also obtained from the same unit.

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† Radio Transmitter Engineering Department, General Electric Co., Schenectady, N. Y.

Initial Considerations

With the inauguration of the first high-frequency frequency-modulation broadcast programs, the need for high-fidelity studio-to-transmitter radio circuits became apparent. The assignment of the frequency band 330.4 to 343.6 megacycles by the Federal Communications Commission to studio-to-transmitter service gave the "all clear" to the development of suitable equipment to serve this need.

The rigid requirements on over-all performance as regards frequency response, distortion, and noise level for frequency-modulation stations indicate that the relay-circuit characteristics must be held to very close tolerances. The Federal Communications Commission requirements for high-frequency broadcast stations[1] include over-all characteristics from microphone input to main transmitter output. Specifically, the noise level must be at least 60 decibels below 100 per cent modulation, the distortion must not exceed 2 per cent

[1] Federal Communication Commission Rules and Regulations, Section 3.245(A). Also, Standards of Good Engineering Practice Concerning High Frequency Broadcast Stations, Section 6, paragraph A-(2, 3, 4).

root-mean-square at 100 per cent modulation, and the frequency response must be within 2 decibels of the standard Radio Manufacturers Association 100-microsecond pre-emphasis characteristic. Distortion and frequency-response measurements must be made over the range 50 to 15000 cycles per second. The 2 per cent distortion requirement has been temporarily relaxed to 3 per cent.[2]

The General Electric type GF-8-A studio-to-transmitter transmitter and model LM-156 studio-to-transmitter receiver which will be described have been designed to meet these requirements.

One question which arose early during the development of this equipment was in regard to what frequency swing should be established as "100 per cent" modulation. While a frequency swing up to ±200 kilocycles is allowed for this service, it appeared more desirable to use ±75-kilocycle swing as a reference as is done on the frequency-modulation broadcast band. By so doing, all measuring equipment, such as frequency and modulation monitors which have already been designed for accurate measurements on the frequency-modulation broadcast frequencies 42 to 50 megacycles at a frequency swing of ±75 kilocycles, are applicable for measurements on studio-to-transmitter stations simply by providing a high-frequency converter to bring the carrier within the usable range of this equipment. Also the problems encountered in keeping receiver distortion at a minimum are greatly reduced when using the lesser swing.

Further, it is seen that if the allowable ±200-kilocycle swing were used, new transmitter and receiver circuits would have to be developed as well as all associated measuring equipment. On the other hand, if the broadcast swing of ±75 kilocycles were used, much less fundamental development work would be required, since proved and successful designs could be applied to the new system. In fact, part of the basic design of the General-Electric type GF-1-B frequency-modulation broadcast exciter[3] was incorporated in this studio-to-transmitter transmitter.

Federal Communications Commission regulations state that for studio-to-transmitter service, only as much power as is required for satisfactory operation shall be used. In addition, rigorous antenna requirements are set up. It is difficult to predetermine just how much power will be required to satisfy the requirements at all installations. However, the General Electric type MY-36-A studio-to-transmitter antenna is designed to give an effective power gain of 10. If the same antenna is used at the receiving location, an over-all antenna power gain of 100 can be realized. For a transmitter carrier power of 25 watts, the signal at the receiver input would be equivalent to that which would be obtained if 2500 watts and simple dipole antennas were used. The 25-watt carrier power of the type GF-8-A transmitter is, therefore, believed to be ample to provide sufficient signal strength over any distance now contemplated for this service.

Transmitter

The transmitter employs the familiar principle of direct-frequency modulation of an oscillator having its

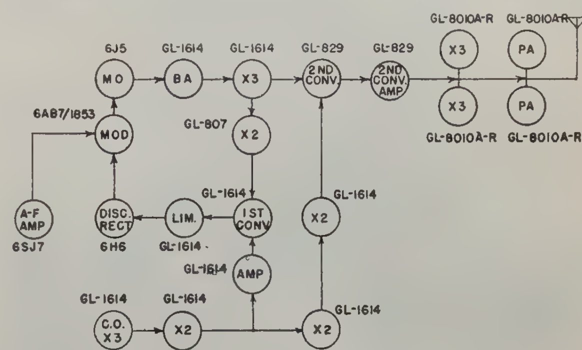

Fig. 1—Block diagram of 25-watt studio-to-transmitter transmitter.

mean frequency stabilized by a crystal (Fig. 1). A frequency-comparison circuit maintains the oscillator on its correct frequency and provides audio feedback resulting in low distortion and noise level. The sixth harmonic of the modulated oscillator and the sixth harmonic of a temperature-controlled, highly stable crystal oscillator are combined to produce an intermediate frequency of 3 megacycles. A temperature-compensated discriminator operating at this frequency produces the direct-current potential for electronic frequency control and also serves as a high-quality demodulator for the audio feedback circuit.

The third harmonic of the modulated oscillator is combined with the twenty-fourth harmonic of the same crystal oscillator used for the frequency-control circuit. These two frequencies, when added together, produce a signal which when tripled becomes the transmitter output frequency.

Since the sixth harmonic of the oscillator is maintained at a frequency 3 megacycles above the sixth harmonic of the crystal oscillator, the frequency of the modulated oscillator is only one-half megacycle above the frequency of the crystal oscillator. Now, if the third harmonic of the modulated oscillator is added to the twenty-fourth harmonic of the crystal oscillator, the resulting frequency is composed of a very large percentage of crystal-channel frequency and a very small percentage of modulated-oscillator-channel frequency. Tripling of this frequency to produce the output or carrier frequency does not change these percentages. The output frequency may be said to be approximately seven eighths directly crystal-controlled and one eighth stabilized master-oscillator controlled.

The modulated oscillator which contributes only approximately one eighth of the output frequency is highly stabilized by the frequency-control circuit. The sixth harmonic of the modulated oscillator and the sixth

[2] As permitted by the Federal Communication Commission Public Notice No. 47160, dated January 31, 1941.

[3] H. P. Thomas and R. H. Williamson, "A commercial 50-kilowatt frequency-modulation broadcast transmitting station," Proc. I.R.E., vol. 29, pp. 537–545; October, 1941.

harmonic of the crystal oscillator are of much higher frequency than the intermediate or difference frequency of 3 megacycles. Since the crystal oscillator is extremely stable, any variations of frequency on the modulated-oscillator channel will result in approximately the same number of cycles variation of the intermediate frequency. But this variation, while probably a very small per cent of the modulated-oscillator frequency, is a very much larger per cent of the intermediate frequency. Thus a great magnification of the modulated-oscillator frequency instability in per cent is obtained for control purposes.

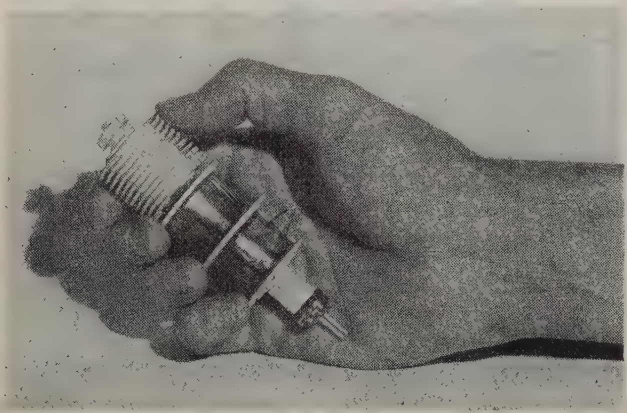

Fig. 2—High-frequency triode, type GL-8010A-R.

This intermediate frequency is then applied to a form of frequency-modulation detector or discriminator, similar to those used in frequency-modulation receivers, which converts frequency changes into amplitude changes. The output voltage of this discriminator will consist of a direct-current component having a magnitude and polarity dependent upon the mean deviation of the modulated-oscillator frequency from its proper value, and audio-frequency components resulting from the frequency modulation of this oscillator. This combined signal is passed through a network which has zero attenuation for the direct-current component and about 20 decibels attenuation for frequencies above 20 cycles. The audio output from this network provides a degenerative signal which is applied to the modulator grid.

The discriminator is so adjusted by extremely stable tuned circuits that when the intermediate frequency is exactly the correct value, no direct output voltage is obtained from the discriminator rectifier. Hence nothing happens to change the modulated-oscillator frequency. If the modulated-oscillator frequency for some reason tends to drift slightly higher, the intermediate frequency changes six times as much and the discriminator rectifier immediately produces a direct voltage of such polarity that, when applied to the grid of the modulator tube, it causes a corrective change to force the oscillator back on its proper frequency. Had the oscillator frequency drifted lower, a direct voltage of opposite polarity from the discriminator would have caused the frequency-modulator tube to correct the oscillator in the opposite direction to that resulting from the previous drift. Thus it is seen that, in addition to the modulated channel contributing only a small part of the output frequency, it is highly stabilized and its frequency depends directly upon the stability of the crystal oscillator and the intermediate-frequency circuits. This results in an over-all stability which is practically as good as that of the crystal alone. Although the required stability is 0.01 per cent (or approximately 34 kilocycles), the guaranteed stability of the output frequency is 0.002 per cent. Measurements during actual operation show that the frequency may be expected to remain within 2 or 3 kilocycles of its assigned value.

A relay is connected to the discriminator in such a way that any failure of the frequency-correction circuit will cause the relay to drop out and shut down the transmitter, thus preventing any possibility of operation of the transmitter without frequency stabilization.

As previously mentioned, the same feedback circuit which performs the frequency-control function also provides feedback for maintaining low noise and distortion levels. A noise level of -70 decibels (unweighted) and a harmonic distortion better than one and one half per cent are maintained.

Audio voltage is supplied to the modulator from a single audio amplifier stage which has an inductance-resistance pre-emphasis network in its plate circuit. The over-all audio-frequency response is within ± 1 decibel of the desired 100-microsecond pre-emphasis standard from 30 to 16,000 cycles.

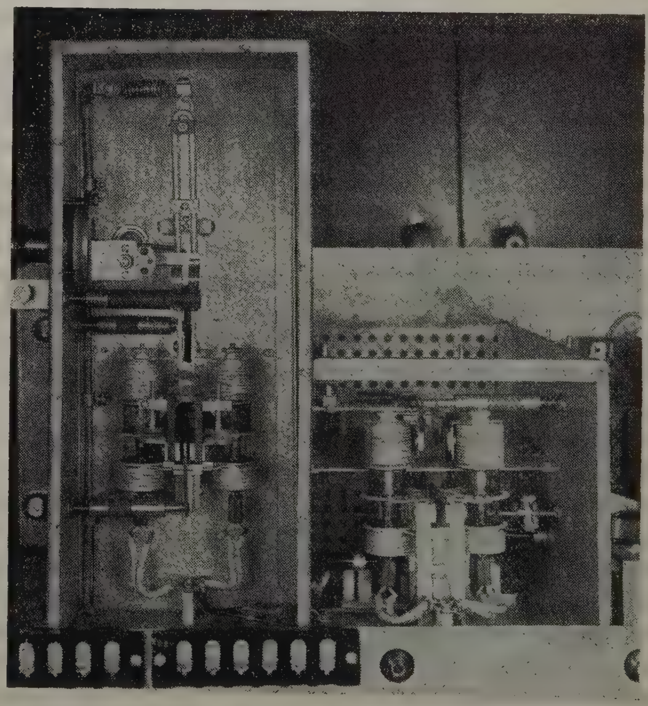

Fig. 3—Transmitter final tripler (right) and 25-watt output stage (left).

The final tripler and output stages utilize type GL-8010A-R tubes (Fig. 2). These newly developed high-frequency triodes have a rated plate dissipation of 50 watts. Full ratings may be used up to 350

megacycles in conventional circuits such as used in the transmitter. The tubes may be used at much higher frequencies when different type circuits are employed.

Both the tripler and output stages have transmission-line tank circuits (Fig. 3). Neutralization is necessary only on the output stage. As a result of careful circuit design, these stages may be adjusted and operated in much the same manner as conventional low-frequency equipment, which is unusual in apparatus operating at ultra-high frequencies.

The radio-frequency output circuit is designed to feed a single coaxial 7/8-inch transmission line having a surge impedance of approximately 65 ohms.

Tube cooling is obtained from small quiet blowers mounted inside the transmitter cabinet (Fig. 4).

Indicating instruments are provided for filament voltage, plate voltage (output stage), and frequency deviation (discriminator direct-current output) (Fig. 5). In addition, three plate milliammeters may be used to check the various circuits by means of transfer switches.

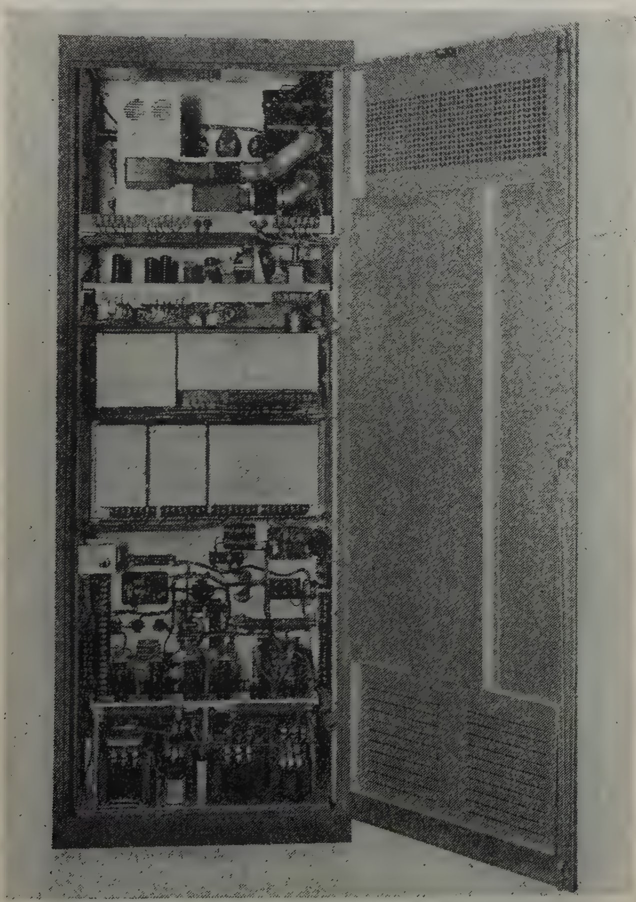

Fig. 4—Rear view of transmitter, door open.

The transmitter operates from a 115-volt, 60-cycle, single-phase power supply. Total power consumption is approximately 1.5 kilowatts.

Access to the interior of the cabinet is gained through full-length doors both front and rear. The front door may be opened for all tuning operations with complete safety to the operators (Fig. 6). Sections of the front door may be opened to gain access to the meter switches and to certain of the tuning controls that are used for the occasional adjustments required to maintain the transmitter at optimum performance.

The transmitter is designed to operate over the range from 260 to 350 megacycles. Thus all the studio-to-transmitter frequencies are covered as well as certain channels for television—sound broadcasting and relaying.

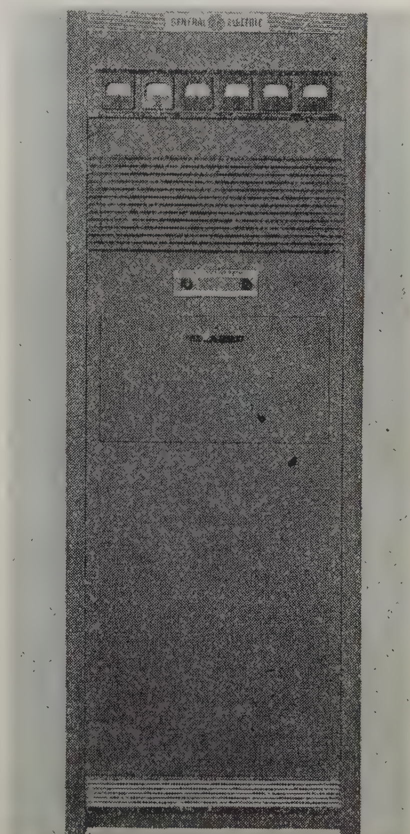

Fig. 5—Front view of transmitter, door closed.

Receiver

The model LM-156 receiver is the companion unit of the type GF-8-A transmitter. Each unit is designed for a particular frequency in the band from 260 to 350 megacycles. This fixed frequency is accurately established by a temperature-controlled low-drift quartz crystal.

The receiver is a double-conversion superheterodyne wherein the oscillator voltage for both conversions is obtained from the same crystal oscillator (Fig. 7). When operating over the band 260 to 300 megaycles, the eighteenth harmonic of the crystal is used for the first conversion and the second harmonic for the second conversion. Over the range from 300 to 350 megacycles, the sixteenth harmonic is used for the first conversion and the second harmonic for the second conversion.

"Acorn"-type tubes are used in the high-frequency stages. A radio-frequency stage precedes the first converter to improve the intermediate-frequency rejection ratio.

All the high-frequency circuits are linear tuned

circuits. Coarse frequency adjustment is accomplished by movable short-circuiting bars and vernier adjustments may be made with small trimmer capacitors. All high-frequency circuits are silver-plated for good conductivity.

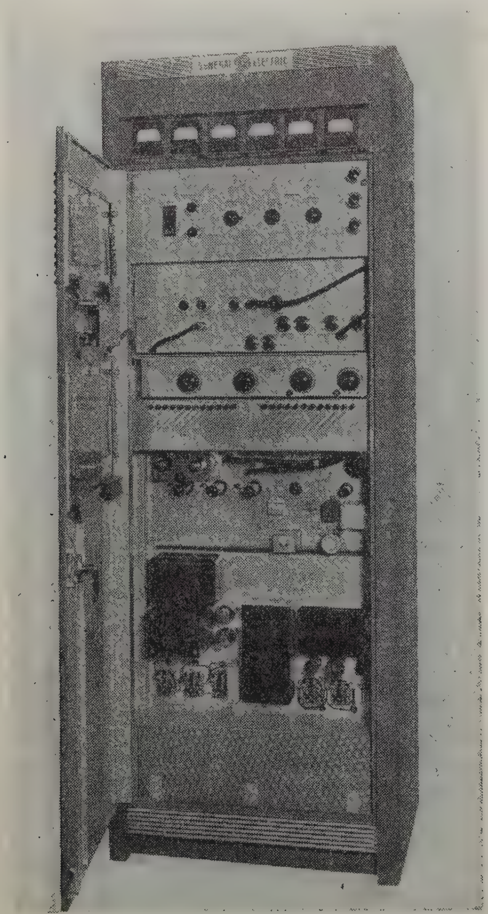

Fig. 6—Front view of transmitter, door open.

All intermediate-frequency circuits except the discriminator use single-control, permeability tuned transformers making alignment extremely simple. Double limiting is used in the 4.3-megacycle second intermediate frequency. A carrier-off-noise-suppressor circuit is provided with adjustable threshold.

Two audio channels are provided; one for program and the other for monitoring. Each channel is equipped with a resistance-capacitance de-emphasis circuit which

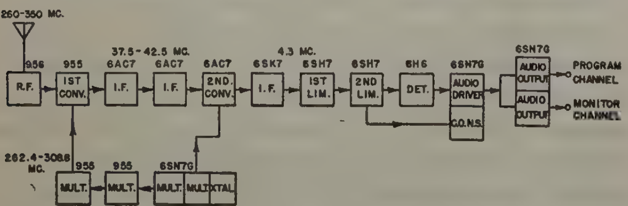

Fig. 7—Block diagram of studio-to-transmitter receiver.

maintains the receiver frequency response within 0.5 decibel of the desired 100-microsecond de-emphasis standard from 30 to 16,000 cycles. Both channels are designed to operate into standard 600-ohm speech-input equipment.

The receiver is equipped with two indicating instruments; one, a high-sensitivity microammeter connected to the discriminator output to indicate detector balance, and the other a milliammeter which may be transferred

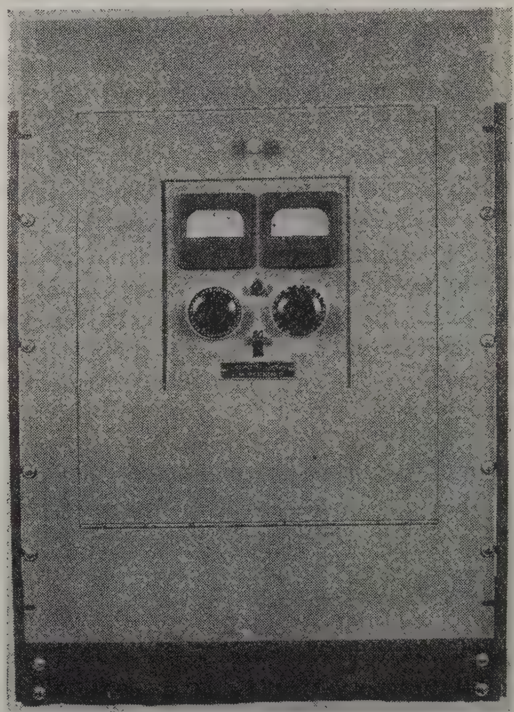

Fig. 8—Studio-to-transmitter receiver.

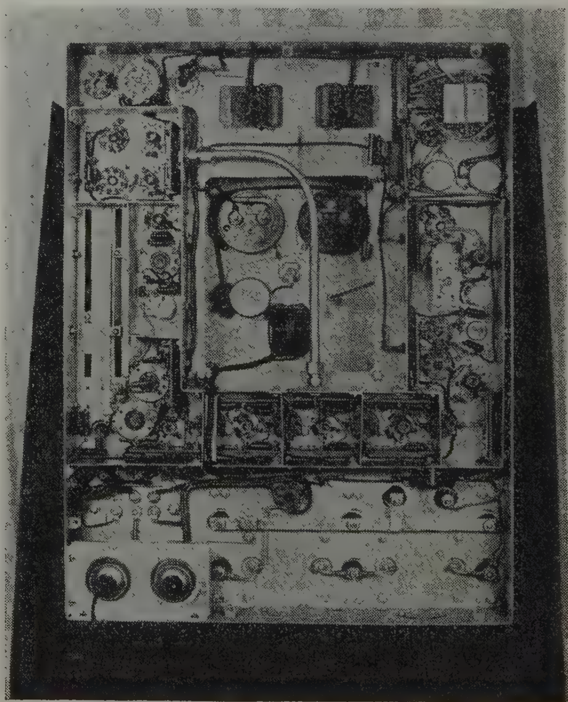

Fig. 9—Rear view of studio-to-transmitter receiver, cover removed.

by means of a transfer switch to the various circuits for tuning adjustments (Fig. 8).

Only one tuning control is brought to the receiver panel. This is the vernier control for the discriminator-secondary tuning capacitor. Normally only occasional

slight adjustment is required since this circuit is temperature-compensated.

Provision is made so that the receiver may be turned on and off with a three-way local-remote switch when it is desired to operate the receiver at a location remote from the operator.

The electrical design of this receiver is such that all circuits are very stable in operation. Shielding is not critical, which is evidenced by the fact that the rear chassis cover may be removed without affecting the operation (Fig. 9). Particular care has been taken to eliminate the possibility of spurious responses when the receiver is operated near other transmitters.

All tuning adjustments are extremely simple. The use of the tuning-meter transfer switch makes it possible to adjust all circuits quickly for optimum operation. The band-pass characteristic is sufficiently wide to pass the frequency-modulated wave without causing distortion.

An important mechanical feature of this receiver is the vertical-chassis construction which provides complete accessibility of all parts. The receiver panel is designed for standard-rack mounting.

All tubes, crystals, tuning controls, and adjustments are accessible through the front-panel door (Fig. 10). The chassis design and arrangement of parts allow complete ventilation.

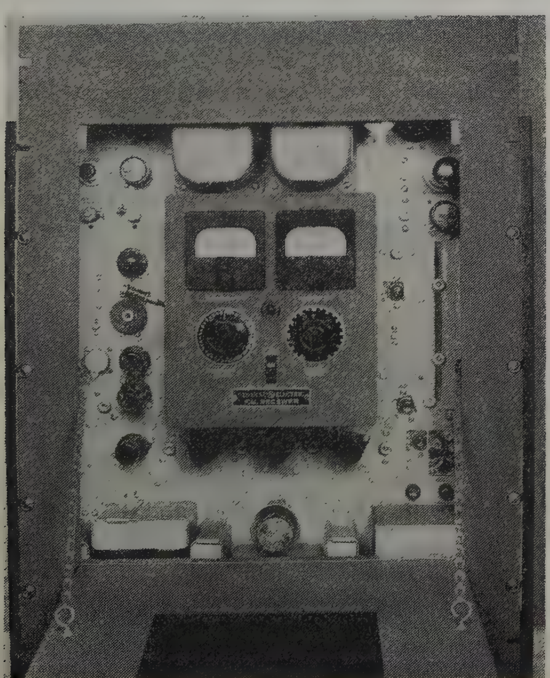

Fig. 10—Studio-to-transmitter receiver with front door open.

The use of electrolytic capacitors has been completely eliminated in the interest of maximum reliability.

Antenna

Present Federal Communications Commission regulations[4] state that the power gain of a studio-to-transmitter transmitting antenna toward the receiver shall be ten times the free-space field from a doublet and in all other directions 30 degrees or more off the line of the receiver the power gain shall not exceed one quarter the free-space field from a doublet. The type MY-36-A studio-to-transmitter antenna has been designed to meet these specifications.

This antenna consists of two horizontally polarized, colinear arrays. Fig. 11 is a drawing showing the mechanical arrangement.

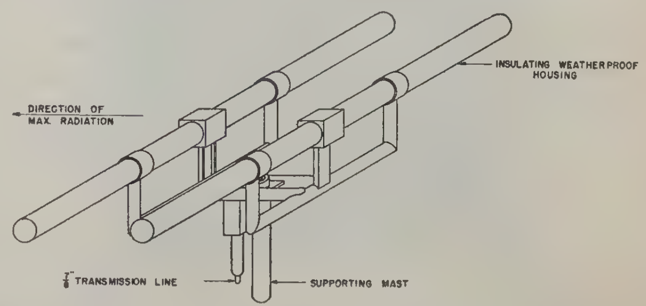

Fig. 11—Outline drawing of studio-to-transmitter antenna.

The studio-to-transmitter antenna is designed to operate over the range of studio-to-transmitter frequencies 330.4 to 343.6 megacycles. No antenna adjustments are required if it is desired to change transmitter frequency within this range.

An important mechanical feature of this antenna is the complete enclosure of the radiating elements, phase inverters, and matching sections within Herkolite insulating tubing. This tubing is airtight and connected so that the entire antenna as well as the transmission line may be pressurized if desired. By enclosing the antenna, all sleet and ice-melting problems are avoided.

Frequency Monitor

It was mentioned above that if a frequency swing of ± 75 kilocycles were established as "100 per cent modulation," standard measuring equipment which has been developed for use in high-frequency frequency-modulation broadcast stations could be modified for use at the studio-to-transmitter frequencies. The use of a high-frequency converter with the General Electric frequency-modulation station monitor[5] (Fig. 12) provides an excellent system for monitoring the studio-to-transmitter output.

This versatile monitor measures four essential characteristics of the frequency-modulated carrier as follows: (1) mean frequency of carrier with and without modulation, (2) percentage of frequency modulation with (3) alarm indication for overmodulation, and (4) fidelity of the modulated signal. In addition, the converter unit which is used with the monitor provides a continuous indication of the transmitter relative output power.

The output of a highly stable, temperature-compensated crystal oscillator is multiplied and heterodyned

[4] Federal Communication Commission Rules and Regulations, Section 4.34(d).

[5] H. R. Summerhayes, Jr., "A frequency-modulation station monitor," Proc. I.R.E., vol. 30, pp. 399–404; September, 1942.

against the transmitter carrier, creating an intermediate-frequency component. This intermediate frequency passes through a current-sensitive discriminator, the output of which actuates the direct-reading frequency

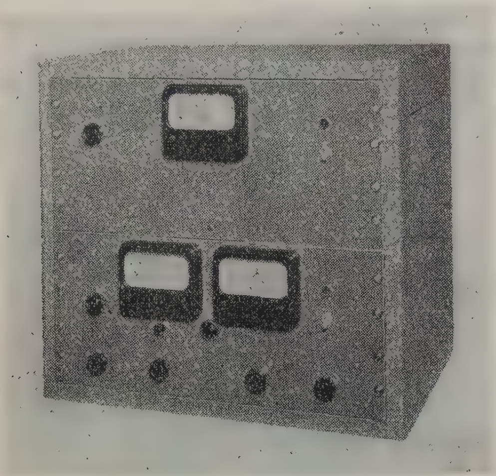

Fig. 12—Frequency-modulation station monitor and high-frequency converter.

meter. A calibration crystal is provided for adjustment of the discriminator circuit.

The discriminator output also actuates the alarm circuit which may be adjusted to operate at any predetermined modulation percentage over a range of 50 to 120 per cent. In addition, an output amplifier is provided which may be used for aural monitoring of the transmitted signal. A conventional VU meter of short

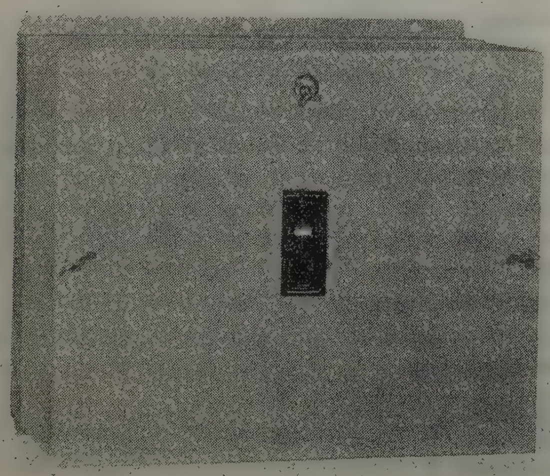

Fig. 13—Transmitter remote-control unit, for use at transmitter location.

time constant is used for continuous indication of modulation percentage.

Remote-Control Units

Remote control of studio-to-transmitter transmitters is not authorized by the Federal Communications Commission. However, if the following conditions are satisfied, the operation does not constitute the prohibited "remote control" of the transmitter: (1) The operator can reach the transmitter in five minutes. (2) The operator has off-on control of power to the last radio-frequency stage. (3) There are continuously indicating instruments at the operating position for frequency deviation and percentage modulation indication; there is also a spurious emission check (aural monitoring).

It is seen that the use of the above monitor converter makes it possible to satisfy all requirements with the exception of the transmitter off-on control. Remote-control equipment for operation of the transmitter from a distant point over a two-wire circuit (such as a telephone line) and ground has been designed for this purpose. This remote-control equipment consists of two units, one located near the transmitter (Fig. 13) and the other at the control point (Fig. 14). These units

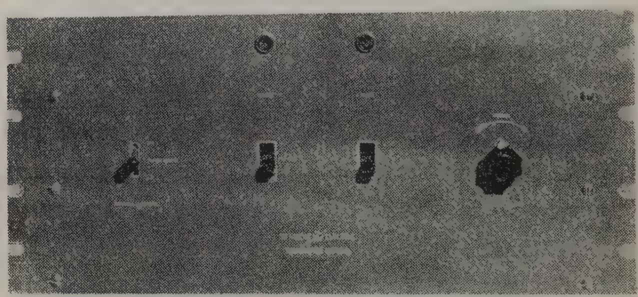

Fig. 14—Transmitter remote-control unit, for use at control point.

contain the necessary switches and relays for placing the transmitter in and out of operation from the remote point without the continued presence of the operator at the transmitter.

Conclusion

The studio-to-transmitter transmitter, receiver, and associated equipment which have been described are commercial types which will meet all the performance requirements for studio-to-transmitter service. The operation and control of the equipment has been kept as simple as possible while still maintaining all the features which are customary in standard broadcast equipment. One fundamental basis for the design of the transmitter was simplicity and ease of maintenance both from a mechanical and electrical point of view.

Several installations have been in successful operation for over one year.

Acknowledgment

It must be recognized that many individuals have contributed to the electrical and mechanical design of this equipment. Acknowledgment should be given particularly to the work of Howard M. Crosby on the transmitter, G. W. Fyler on the receiver, M. W. Scheldorf on the antenna, and H. R. Summerhayes, Jr., on the monitor, all of the General Electric Company.

Power-Tube Performance in Class C Amplifiers and Frequency Multipliers as Influenced by Harmonic Voltage*

ROBERT I. SARBACHER†, MEMBER, I.R.E.

Summary—**In investigating the effect of harmonic voltage in class C amplifiers, it is found that the second- or third-harmonic voltages introduced into the plate or grid circuit in the proper phase improve the performance of the tube. Harmonics of higher order are found to be undesirable in general. The most favorable path of operation is determined and it is found that this path may be easily obtained by the introduction of the third-harmonic voltage 180 degrees out of phase with the fundamental voltage into the plate circuit. Various means for doing this are devised, and the power output and over-all efficiency of the tube, acting as a class C amplifier, is increased. The method of analysis employed, when applied to frequency multipliers, shows that by the introduction of harmonic voltages into these devices in the proper phase and magnitude, substantial increases in power output and efficiency may be obtained.**

Ever since their introduction into the field of communication engineering over twenty-five years ago, high-vacuum electron tubes of the power type have assumed a steadily increasing importance. Improvement in design technique has permitted the construction of tubes giving outputs of thousands of watts. Owing to the use of such large power, any factors which contribute to an increase in the efficiency of operation of the electron tubes are worthy of serious consideration. Many papers have appeared which deal with technical investigations of the properties of tubes operating as radio-frequency class C amplifiers. These papers are concerned largely with the conditions under which optimum conversion of direct-current plate-supply power into radio-frequency power is obtained, consistent with the demands of the type of service required. One of the factors governing the efficiency in power amplifiers is concerned with the introduction of harmonic voltages into the amplifying system, and this factor has, up to this time, received little attention.

In an attempt to analyze this problem it has been found desirable to investigate the most favorable path of operation for power-tube performance. Let us consider first the most favorable path of operation for the plate circuit; that is, the one which gives the maximum plate-circuit efficiency for a given power output. The plate efficiency may be expressed as

$$(\mathrm{eff})_p = (E_{p_1m}I_{p_1m})/(2E_{bb}I_{ba})$$

where E_{p_1m} = the maximum value of the fundamental component of plate voltage

I_{p_1m} = the maximum value of the fundamental component of plate current

E_{bb} = the plate-polarizing potential

I_{ba} = the average value of plate current

If we fix, for the moment, the ratio E_{p_1m}/E_{bb}, the efficiency becomes a function of the ratio I_{p_1m}/I_{ba}. The problem is to determine the path joining the quiescent point Q, Fig. 1, with the point A so as to make this ratio a maximum. Now

$$\frac{I_{p_1m}}{I_{ba}} = \frac{2\int_0^{\theta_1}[I_b(e_b, e_c)]\cos\theta d\theta}{\int_0^{\theta_1}[I_b(e_b, e_c)]d\theta}$$

where $I_b(e_b, e_c)$ represents the equation of the path of operation on the static plate-current characteristic surface of the tube, and θ_1 is the half angle during which plate current flows.

It can be seen by inspection that as $\theta \rightarrow 0$, the $\cos\theta \rightarrow 1$ and the ratio I_{p_1m}/I_{ba} approaches its maximum value 2. It is evident that θ_1 must be a minimum for maximum efficiency when E_{p_1m}/E_{bb} is fixed. Now if $E_{b_1m} \doteq E_{bb}$, the plate efficiency approaches 100 per cent as θ_1 approaches zero. Hence, the most favorable path of operation is one which will penetrate the plate-current region as late in the cycle as possible and rise as quickly as possible to the maximum current while keeping $E_{b_1m} \geqq E_{bb}$ during the interval in which current flows as shown by the lower dashed line in Fig. 1. In order to maintain a fixed power output, it is necessary that the product $E_{b_1m}\, I_{p_1m}$ be maintained constant. As θ_1 is reduced, I_{p_1m} is reduced and hence E_{p_1m} must be increased proportionally. It will be shown that it is possible to do this without increasing the plate-polarizing potential.

From the standpoint of the grid circuit, it is desirable to have the driving power P_d as small as possible. Since this power supplies both the power delivered to the grid P_g, as well as the power delivered to the grid battery or bias supply P_c, it is desirable to keep both of these as low as possible. If the grid-polarizing potential and amplitude of grid-excitation voltage are determined for any given tube, the path that will make P_d a minimum will be the path that will make the fundamental component of the grid current I_{g_1m} and the average value of the grid current I_{ca} a minimum. From considerations analogous to those made in the plate circuit above, we can see that this path, when in the region during which grid current flows, should be orthogonal to the lines of constant grid current as shown by the upper dashed line of Fig. 1.

It is apparent that the path resulting in maximum plate efficiency differs greatly from that most favorable

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† Formerly Cruft Laboratory, Harvard University, Cambridge, Mass., later, Illinois Institute of Technology, Chicago, Ill.; now, Bureau of Ships, Navy Department, Washington, D. C.

to the grid circuit. Indeed each "ideal path" results in exceedingly inefficient operation for the other element. It might appear upon inspection that the normal straight-line path represents a good compromise between these two. It is found, however, that this is not the case, and that a more desirable path is that indicated by the dot-dash line in Fig. 1. For this path the driving power is increased but slightly over that resulting from the ideal grid path, and the plate efficiency still approaches its ideal value without seriously reducing the

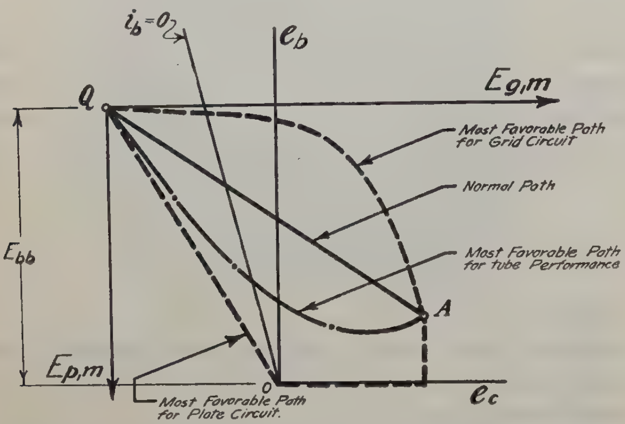


Fig. 1—Paths of operation shown on the e_b-e_c diagram of a power tube. The region in which plate current flows is to the right of the line $i_b=0$ and above the e_c axis. Only the "power stroke" of the path of operation is indicated in the figure.

power outut. The resulting tube efficiency is then considerably greater than it was with the simple straight-line path. Whether or not this advantage is offset by the expense of producing this path of operation, and how closely it may be approximated remains to be shown.

The possibility of obtaining this path of operation by the introduction of harmonic voltages in the grid circuit will be considered first. Fig. 2a shows the effect on the path of operation of the introduction of the second harmonic voltage in the grid. If $e_b=E_{bb}-E_{p_1m}\cos\theta$ and $e_c=-E_{cc}+E_{g_1m}\cos\theta+E_{g_2m}\cos 2\theta$ we shall obtain the paths indicated in this figure. For path a, $E_{g_2m}=25$ per cent E_{g_1m} and for path b, $E_{g_2m}=50$ per cent E_{g_1m}. If the phase of the second harmonic is different from that shown, the operation is less efficient.

Other examples of the shape of the path of operation resulting from the introduction of harmonic voltages into the grid circuit are shown in Fig. 2. Fig. 2b shows the effect of the third harmonic, Fig. 2c the fourth, and Fig. 2d the fifth. The best approximations to the ideal path of operation are obtained by the use of either the second or third harmonic. The fourth harmonic bends the path of operation back into the region of plate current at high plate voltages. If E_{cc} is increased to avoid this, then E_{g_1m} and E_{g_4m} must be increased in order to maintain the same operating point. This will result in an increase in the value of grid-polarizing potential required and also increased driving power. Higher-order harmonics have an even less favorable effect upon the path of operation.

The effect produced by the introduction of harmonic voltages into the plate circuit will now be considered. In Fig. 3 paths a and b show the correct phase relation for the second harmonic. Here $e_c=-E_{cc}+E_{g_1m}\cos\theta$ and $e_b=E_{bb}-E_{p_1m}\cos\theta+E_{p_2m}\cos 2\theta$. In path a, $E_{p_2m}=50$ per cent E_{p_1m}; in b, $E_{p_2m}=25$ per cent E_{p_1m}. When the phase of the second-harmonic voltage is reversed so that $e_b=E_{bb}-E_{p_1m}\cos\theta+E_{p_2m}\cos 2(\theta+(\pi/2))$, path c results. Path d is the normal straight-line path. Figs. 4a to d show the effects of the third, fourth, fifth, and sixth harmonics introduced into the plate circuit. As in the case of the grid circuit we can see from these figures that the harmonics of higher order than the third are of small advantage in realizing the ideal path.

In Fig. 5 the paths obtained with the second and third harmonics in the plate circuit are shown in more direct relation to the static characteristic curves. Both of these paths are excellent approximations to the ideal path. In each case the quiescent point Q has been adjusted for most favorable operation. The use of the second harmonic reduces the necessary grid-polarizing potential; whereas the use of the third harmonic increases it. On the other hand, a relatively large second-harmonic amplitude is required to approach the ideal path, while a comparatively small third-harmonic amplitude is sufficient. The use of increased fundamental voltage over that required for normal operation aids in compensating for the reduction in the amplitude of the fundamental component of plate current caused by the reduced operating angle. This results in approximately constant and sometimes increased power output, even though I_{p_1m} is reduced.

It is possible to introduce these harmonic voltages into the plate or grid circuits without simultaneous introduction of appreciable reactance. The conditions that must be met in order to do this may be stated as follows. If X_{p_k} and X_{g_k} are reactances at the kth harmonic frequency, existing in the plate and grid circuits, respectively, then it is necessary that

$$X_{p_1}\ll(R_b)_{\omega_1} \qquad X_{g_1}\ll(R_c)_{\omega_1}$$
$$X_{p_k}\ll(R_b)_{\omega_k} \qquad X_{g_k}\ll(R_c)_{\omega_k}$$

where $(R_b)_{\omega_k}$ and $(R_c)_{\omega_k}$ are the equivalent resistances of parallel resonant circuits inserted in the plate or grid circuits and tuned to the frequency of the harmonic voltage it is desired to obtain. If, for example, a parallel resonant circuit is inserted in the plate circuit of an amplifier and tuned to the second harmonic, a second-harmonic voltage will be developed across this harmonic tank. When the reactance of the plate circuit is negligible, this harmonic voltage E_{p_km} will be in phase with the harmonic current I_{p_km}. Whether this harmonic voltage will have the correct phase relation to the fundamental voltage to give us the path of operation we desire remains to be determined.

Since the current pulse generated in this type of operation is symmetrical and periodic, Chaffee's simplified harmonic analysis[1] may be used to analyze it. According

[1] E. L. Chaffee, "A simplified harmonic analysis," *Rev. Sci. Instr.*, vol. 7, p. 38; 1936.

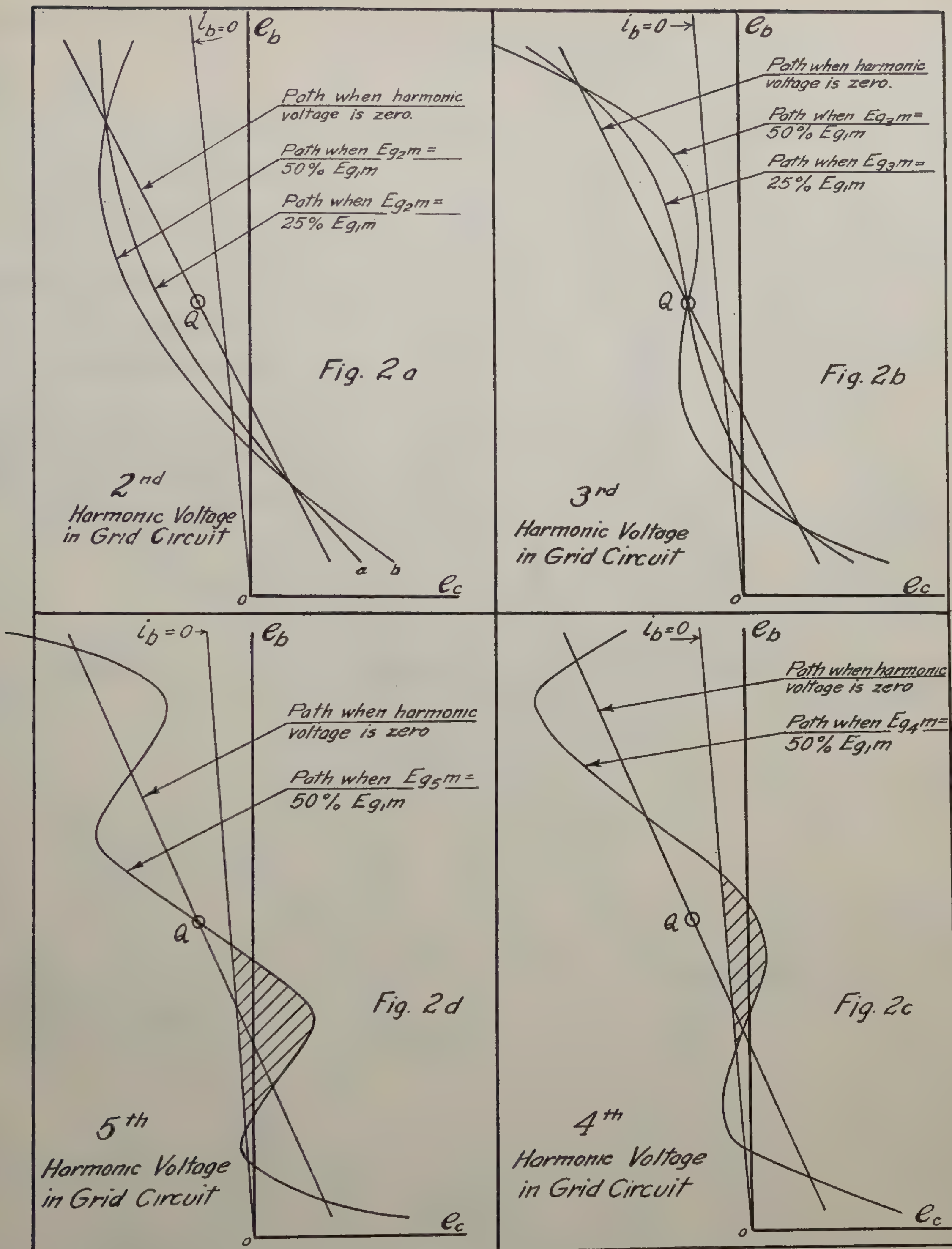


Fig. 2—The path of operation on the e_b, e_c plane of a power tube as influenced by the second-, third-, fourth-, and fifth-harmonic voltages introduced into the grid circuit.

to the "13-point" analysis, further simplified as is allowable with the special type of current pulses considered here, the components of the current wave may be expressed in terms of the instantaneous values. These values are chosen at 15-degree intervals along the time axis of the wave. If C_1 is the value of the instantaneous current at $\theta=0$, C_2 its value at $\theta=15$ degrees, C_3 its value at $\theta=30$ degrees, etc., the analysis tells us that the average and harmonic currents can be defined in terms of the values at these intervals by the schedule given below.

$$I_{ba}=\tfrac{1}{12}(C_1/2+C_2+C_3+C_4+C_5+C_6)$$
$$I_{p1m}=\tfrac{1}{12}(C_1+1.93C_2+1.73C_3+1.41C_4+C_5+0.26C_6)$$
$$I_{p2m}=\tfrac{1}{12}(C_1+1.73C_2+C_3-C_5-1.73C_6)$$
$$I_{p3m}=\tfrac{1}{12}(C_1+1.41C_2-1.41C_4-2C_5-1.41C_6)$$
$$I_{p4m}=\tfrac{1}{12}(C_1+C_2-C_3-2C_4-C_5+C_6)$$
$$I_{p5m}=\tfrac{1}{12}(C_1+0.518C_2-1.73C_3-1.41C_4+C_5+1.93C_6)$$

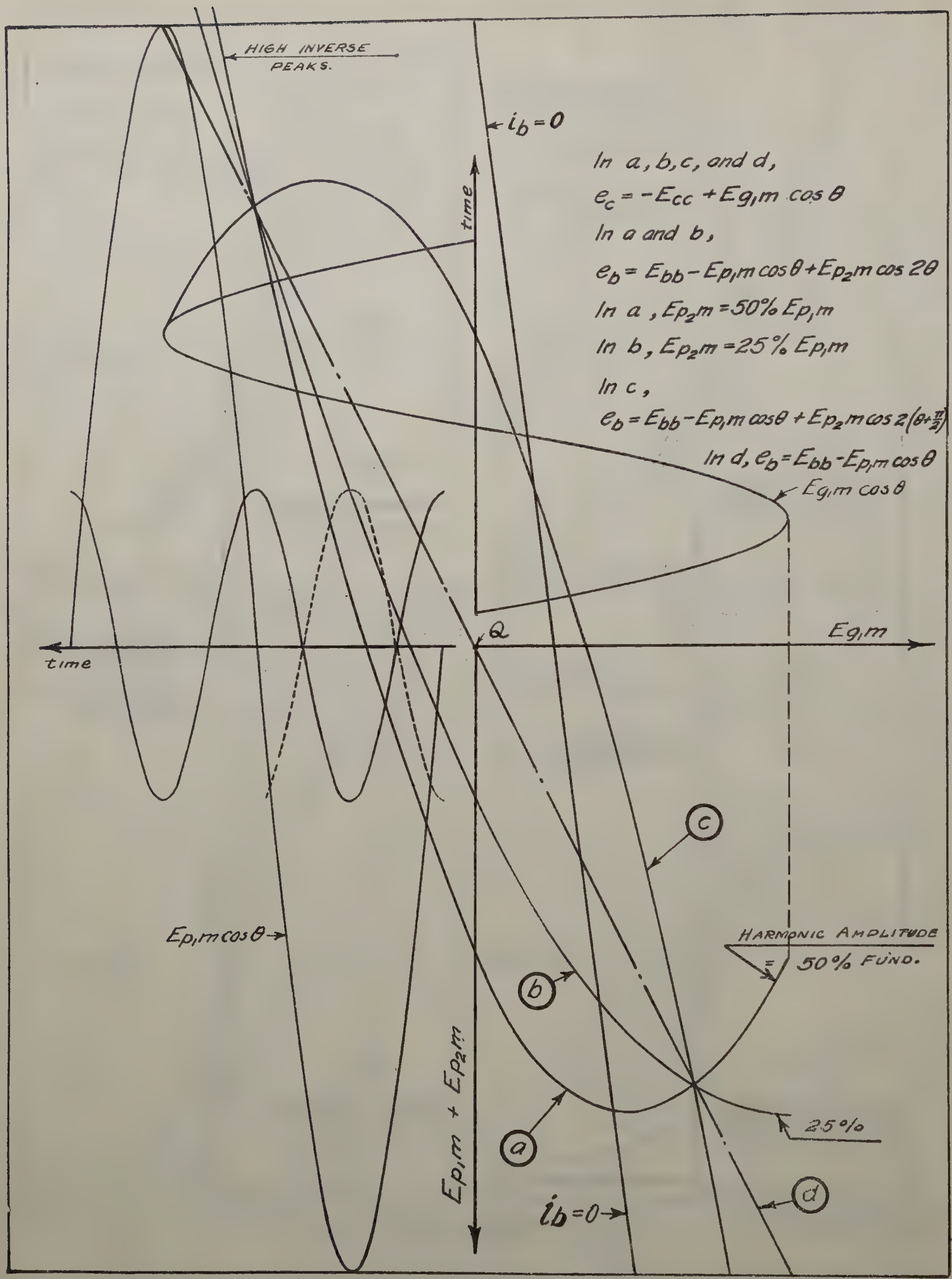


Fig. 3—Path of operation when the second-harmonic voltage is present in the fundamental voltage of the plate circuit.

$$I_{p_6m} = \tfrac{1}{12}(C_1 - 2C_3 + 2C_5)$$

where I_{ba} = the direct-current component of the current wave

I_{p_1m} = the maximum value of the fundamental component in the current wave

I_{p_km} = the maximum value of the kth harmonic component in the current wave

We can now determine the phase relations between the fundamental current in the plate or grid circuits and the harmonic currents in these circuits. It was shown

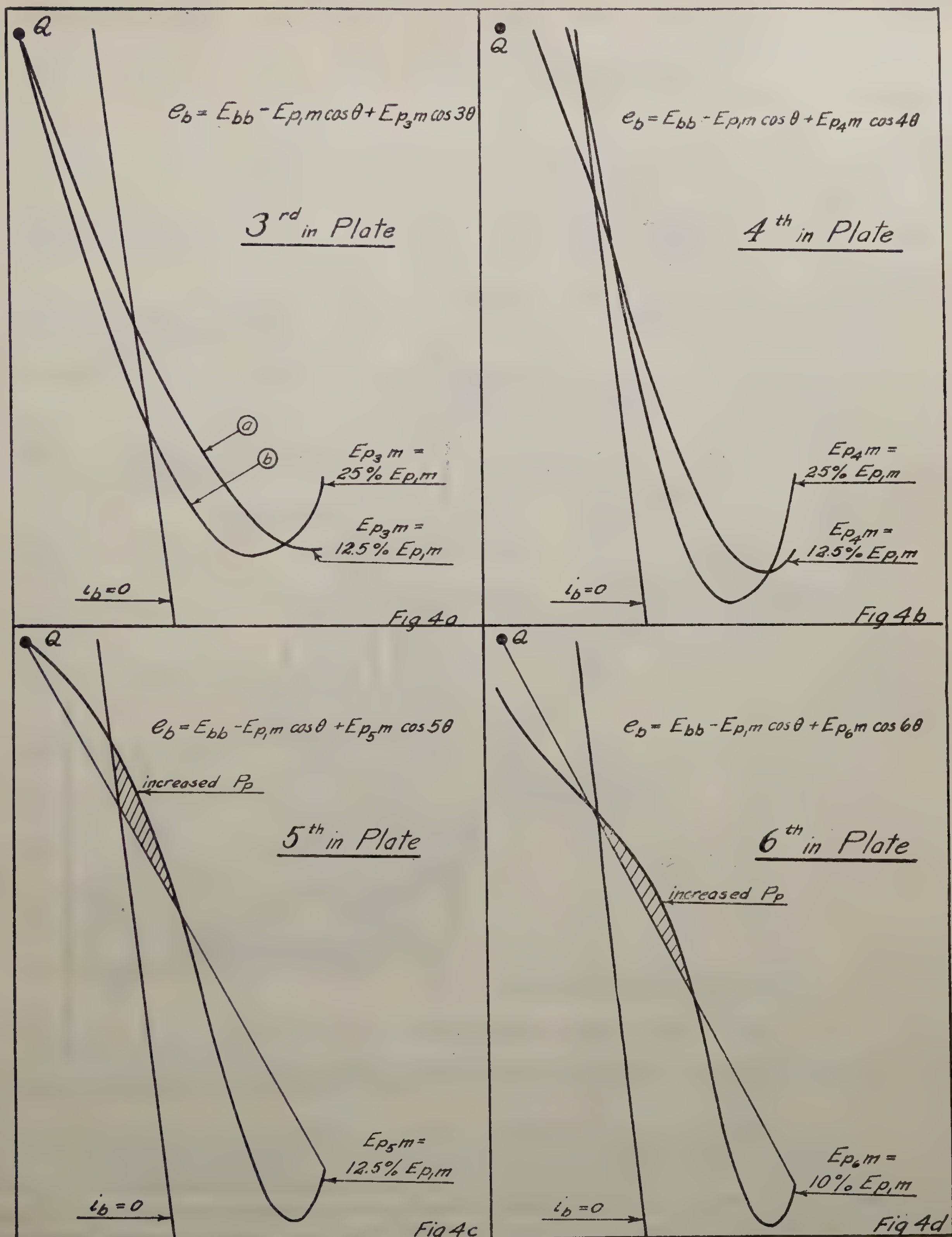


Fig. 4—The path of operation on the e_b, e_c plane of a power tube as influenced by the third-, fourth-, fifth-, and sixth-harmonic voltages introduced into the grid circuit. Only the path during the "power stroke" is shown.

that in the case of the introduction of the second-harmonic voltage in the plate circuit, we obtain the most favorable path when $e_b = E_{bb} - E_{p_1m}\cos\theta + E_{p_2m}\cos 2\theta$. That is, the amplitude of the second-harmonic voltage is opposite in sign to that of the fundamental during the power stroke. To meet this phase requirement, I_{p_2m} must be negative, and, consequently, $C_5 + 1.73\ C_6 > C_1 + 1.73\ C_2 + C_3$. But this is never true even with a straight-line path and is contrary to the conditions that are specified for the most favorable path, namely, that

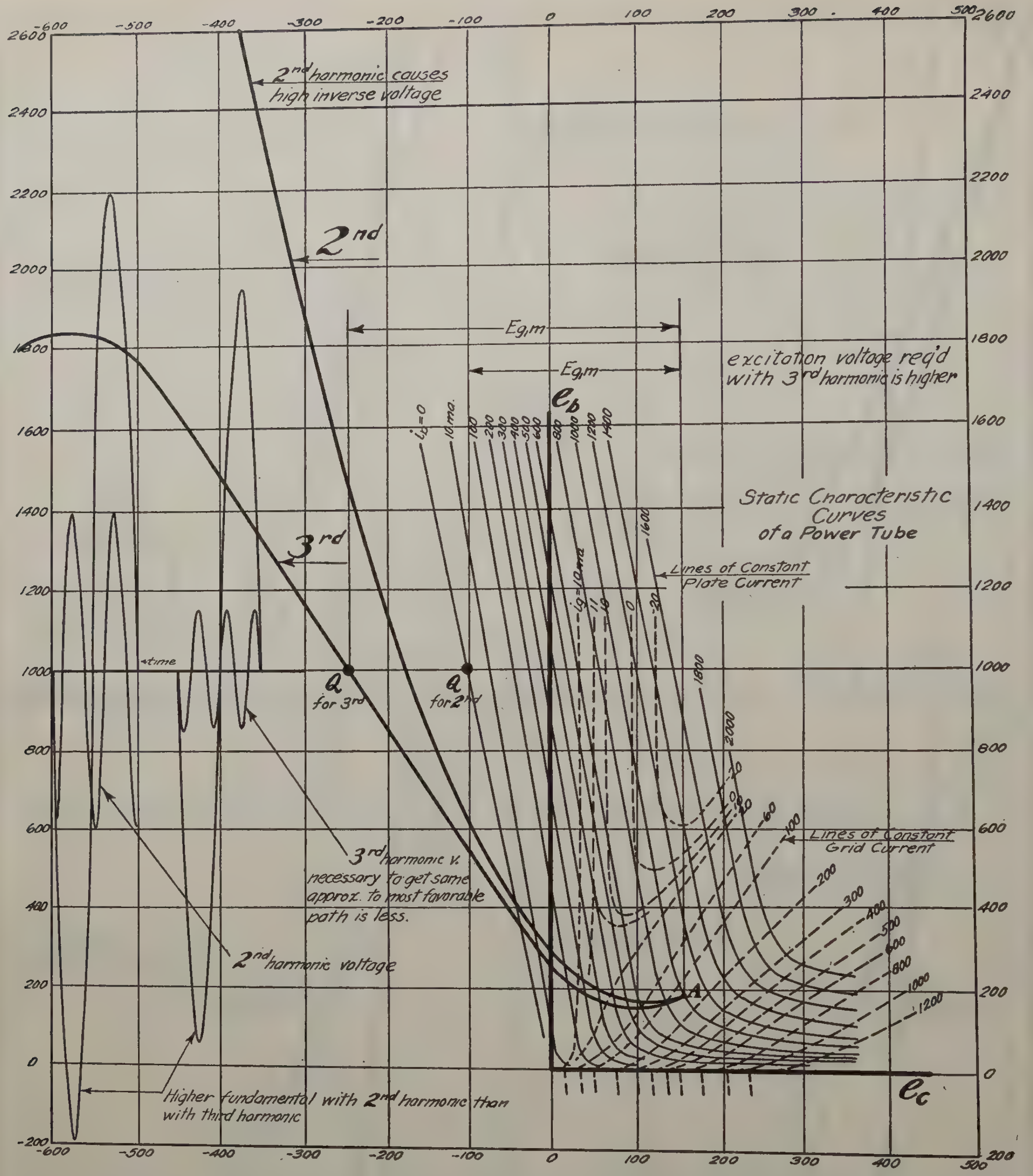


Fig. 5—Paths of operation on the e_b, e_c diagram when the second- and third-harmonics voltages are introduced, in the correct phase, in the plate circuit of a power tube. The end point A of the paths of operation is chosen to be the same for both paths.

the angle during which current flows must be small.[2] Therefore, it appears that the second-harmonic component of the current pulse from the tube produces a second-harmonic plate voltage which is opposite in phase to that required to give the most favorable path of operation and second-harmonic power of the correct phase must be supplied from an external source to buck out this undesired component.

From the harmonic analysis we can see clearly the effect of a variation of θ on the ratio of I_{p_1m}/I_{ba}. From this analysis we obtain

[2] The existence of an instantaneous current at C_6 means that $2\theta > 150$ degrees.

$$\frac{I_{p_1m}}{I_{ba}} = \frac{C_1+1.93C_2+1.73C_3+1.41C_4+C_5+0.26C_6}{C_1/2+C_2+C_3+C_4+C_5+C_6}.$$

If $C_2+C_3+C_4+C_5+C_6=0$, we get the value 2 for this ratio, which we obtained previously. For small increases in θ this ratio is only slightly affected. Now for large power output it is desired to have a high I_{p_1m}, and from this standpoint, the existence of all the C's is desirable. But, since I_{p_1m}/I_{ba} should be high from the standpoint of efficiency, it is desirable to have the terms C_5 and C_6 equal to zero, since their contribution to I_{p_1m} is small. This will limit $2\theta \leqq 120$ degrees. In the case of the third-harmonic voltage introduced in the plate, it is necessary that $1.41\ C_4+2C_5+1.41\ C_6 > C_1+1.41\ C_2$ in order that

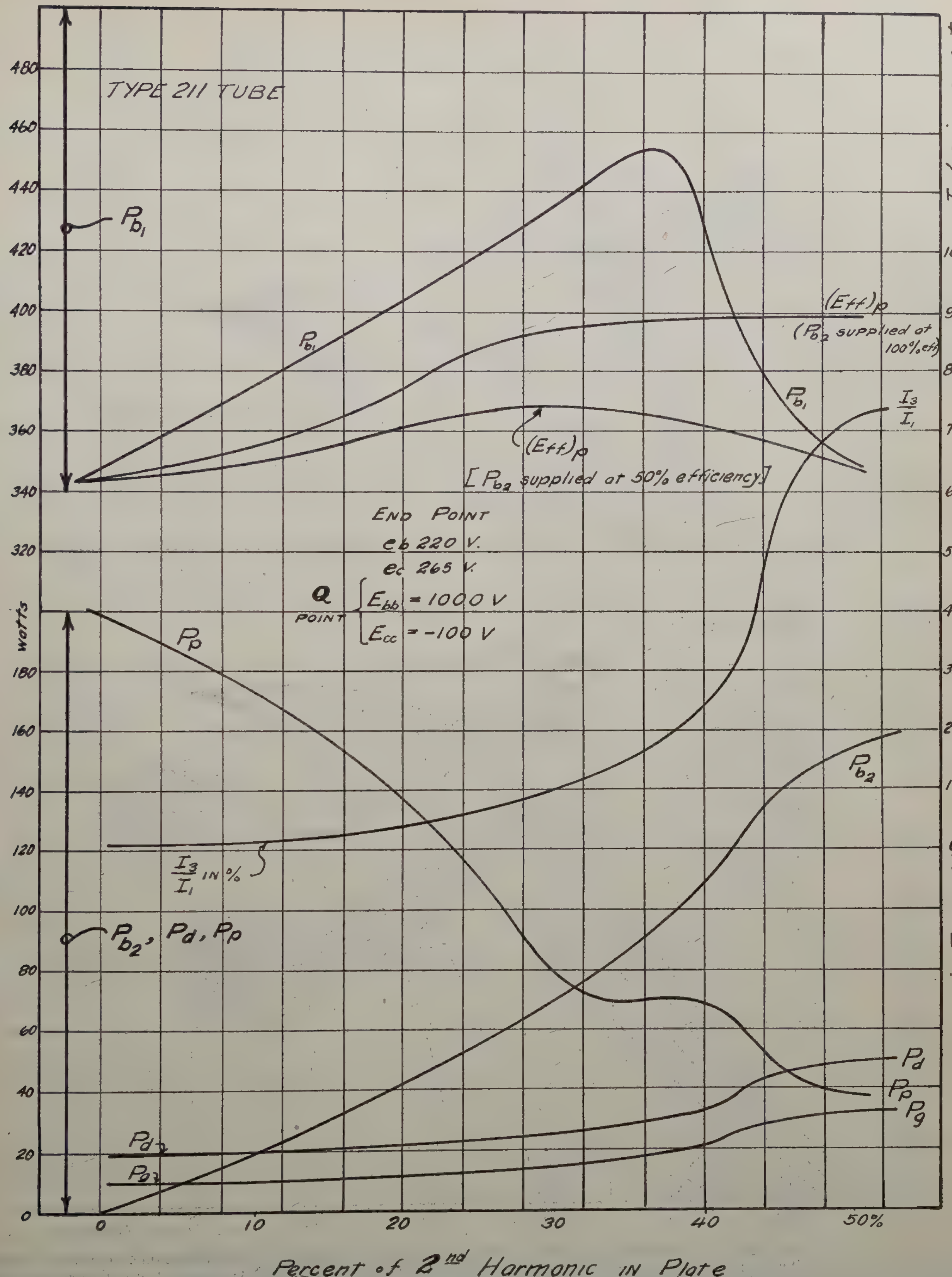


Fig. 6—The variation of the dependent variables of the amplifying system as a function of the second-harmonic voltage injected in the correct phase, into the plate circuit. The second-harmonic voltage is expressed as a percentage of the fundamental voltage amplitude.

the harmonic current develop the voltage in the phase we wish to have it. Since the ideal path requires that θ be small, this condition cannot be met except at a sacrifice in efficiency,[8] and, hence, the third-harmonic power must be supplied in the correct phase from an external source.

[8] There is one exception to this statement which will be discussed later.

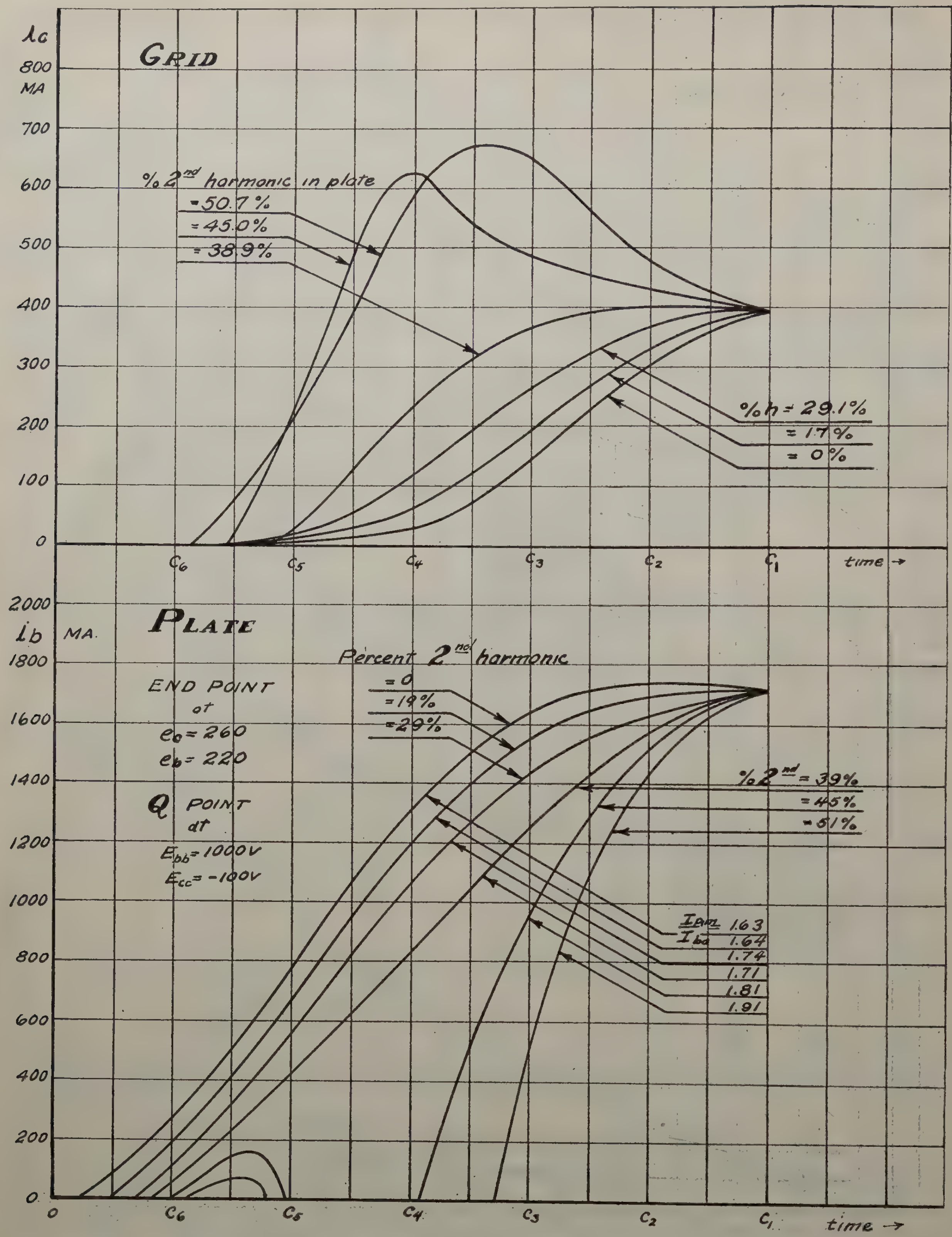


Fig. 7—Instantaneous current pulses in the plate and grid circuits for various percentages of second-harmonic content in the fundamental plate voltage. The quiescent point Q is located at $E_{bb}=1000$ volts, $E_{cc}=-100$ volts and the end point of the path of operation, on the power stroke, at $e_b=220$ volts, $e_c=260$ volts, Fig. 5. Only half of the current pulses are shown. The pulses are symmetrical about the time $t=C_1$ on the diagram.

From similar consideration in the grid circuit it can be shown that the voltage of harmonic frequency developed across a second- or third-harmonic tank inserted in this circuit will be in the wrong direction to give the path of operation desired. Hence, the harmonic power required to maintain the correct voltage must be supplied by an external circuit. The improvement in operation that is obtained in this case is relatively small.

If the power required to maintain a desired harmonic voltage is introduced into the plate circuit, then the plate-circuit efficiency must be expressed as

$$(\text{eff}_p) = P_{b_1}/(P_{bb} + P_{bk})$$

when P_{b_1} = the power output of fundamental frequency
P_{bb} = the direct-current power supplies to the plate circuit
P_{bk} = the power of kth harmonic frequency introduced

In amplifiers which are supplied in this way with harmonic voltage to improve the performance of the tube, the plate efficiency as defined above may be misleading. In standard-frequency doublers, plate efficiencies of 60 to 65 per cent are considered good. For frequency triplers this efficiency may fall as low as 50 to 55 per cent. Hence, the losses incurred in obtaining the harmonic power required should be taken into consideration in making a comparison of the performance of the tube operating in this way, with normal operation.

Because of the higher efficiency of oscillators, it would be desirable to use one to obtain this harmonic voltage. However, this is difficult except in the special case discussed later, since the oscillator will tend to lock in phase with the harmonic current in the opposite phase to that desired.

Fig. 6 shows the effect produced on the dependent variables of the amplifier system, when the second harmonic is introduced in the plate circuit in the correct phase while Fig. 7 shows the shape of the current pulses for different percentages of harmonic voltage with the same constant values of the independent variables.

The plate efficiency can be seen in Fig. 6 to increase with harmonic content, when P_{b_2} is supplied at 100 per cent efficiency. If, however, P_{b_2} is supplied at 50 per cent efficiency, the amplifier efficiency reaches a maximum with approximately 30 per cent second-harmonic voltage. This, of course, is due to the increase in P_{b_2} with increasing harmonic content, while the power output is reduced rapidly as the angle during which current flows is reduced to below 90 degrees. The driving power and grid loss are seen to increase slightly with increased harmonic content, while the plate loss is rapidly reduced. At a second-harmonic content equal to 35 per cent of the fundamental voltage, with 15 watts grid dissipation, 27 watts driving power and 70 watts plate loss, a power output[4] of 450 watts may be obtained. If P_{b_2} is supplied to the system at 50 per cent efficiency, the plate efficiency of the amplifier will be approximately 73 per cent. Chaffee has obtained curves showing the optimum power output that it is possible to obtain with this tube operating under ideal conditions with the straight-line path of operation. These curves show that the optimum power output with the same driving power and plate-polarizing potential, as used above, but with 100 watts plate loss, is slightly under 300 watts. Since P_p = 100 watts, the plate efficiency is slightly under 75 per cent.

From this example, we see that if we can supply P_{b_2} at 50 per cent efficiency, we can obtain approximately 50 per cent more power output with the curved path than we can obtain with the straight-line path, without reducing the plate efficiency to any extent. It will be shown later that, by inserting harmonic voltages into frequency multipliers, large increases in efficiency of these devices may be obtained. Second-harmonic power may then be generated at from 70 to 80 per cent plate efficiency, and third-harmonic power at from 60 to 65 per cent plate efficiency.

Figs. 8 and 9 are similar to Figs. 6 and 7, except that percentages of the third harmonic are plotted as absissas. These diagrams do not represent optimum conditions of operation but are merely representative of the way in which the dependent variables change when the path of operation is varied for any given quiescent point Q and operating point A.

With the harmonic voltage introduced in the plate circuit there are five variables which we may consider as independent. These are E_{bb}, E_{cc}, E_{g_1m}, E_{p_1m}, and E_{p_km}. With such a large number of independent variables, it is convenient to use contour diagrams[5] which show the behavior of the dependent variables, in terms of any two of the independent variables, the other three independent variables being held fixed. Such a diagram is shown in Fig. 10, where E_{bb} and E_{cc} and the ratio $E_{p_3m}/E_{p_1m} = 0.25$ are held fixed and the dependent variables in the system plotted as functions of E_{p_1m} and E_{g_1m} or as functions of e_b and e_c.

In the case of the third-harmonic voltage inserted in the plate circuit, if the independent variables in the amplifying system are adjusted so that $C_6 = 0$ and $1.41\ C_4 + 2C_5$ is just slightly greater than $C_1 + 1.41\ C_2$, a small third-harmonic current will be generated in the plate circuit which will be in the proper phase to develop the desired voltage correctly. This current may then be used to control the frequency of an auxiliary oscillator which will supply the energy necessary to maintain the desired voltage amplitude. Since the losses in a properly constructed parallel resonant circuit may be kept quite low[6] the auxiliary oscillator may be of very low power. When $1.41\ C_4 + 2C_5$ is greater than $C_1 + 1.41\ C_2$ by an amount sufficient to develop enough third-harmonic power to supply the tank losses and maintain the required harmonic voltage, the angle during which plate current flows is usually increased to the point where the efficiency of the system is only slightly improved. Whereas, with the auxiliary oscillator high

[4] Exceptionally high power outputs such as this are of course accounted for by the fact that when the harmonic voltages are introduced into the system in their correct phase relations, the plate dissipation is reduced. This permits us to shift the end point of the path of operation to a region of higher plate current, which causes the plate dissipation and power output to rise. In doing this, care must be taken that the maximum instantaneous plate current does not rise to such a value that the filament life is materially shortened.

[5] E. L. Chaffee and C. N. Kimball, *Jour. Frank. Inst.*, vol. 221, p. 237; 1936.

[6] With the higher frequencies, tuned concentric cables or cavity resonators may be used with which the selectivity Q may exceed 1000.

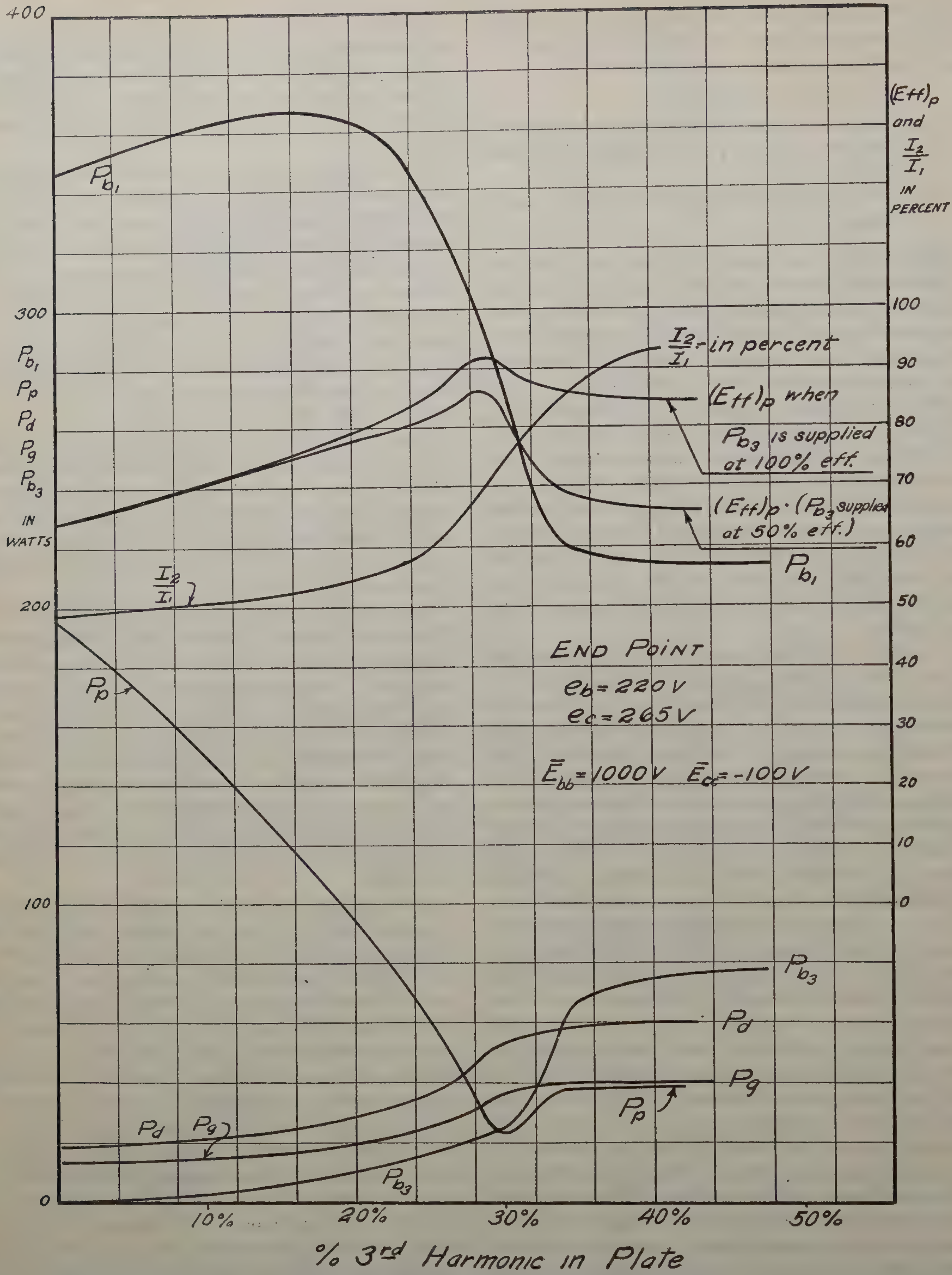


Fig. 8—The variation of the dependent variables of the amplifying system as a function of the third-harmonic voltage, injected in the correct phase, into the plate circuit.

efficiency and high power output can be maintained simultaneously with $C_6=0$ and C_5 very nearly 0; e.g., $2\theta \doteq 120$ degrees.

Figs. 11 and 12 show the dynamic characteristics, taken at a radio frequency of 3.5 megacycles for a type HK-54 Gammatron power tube. The values of the independent variables E_{g_3m} and E_{p_3m} given on these figures represent very nearly optimum values for the plate

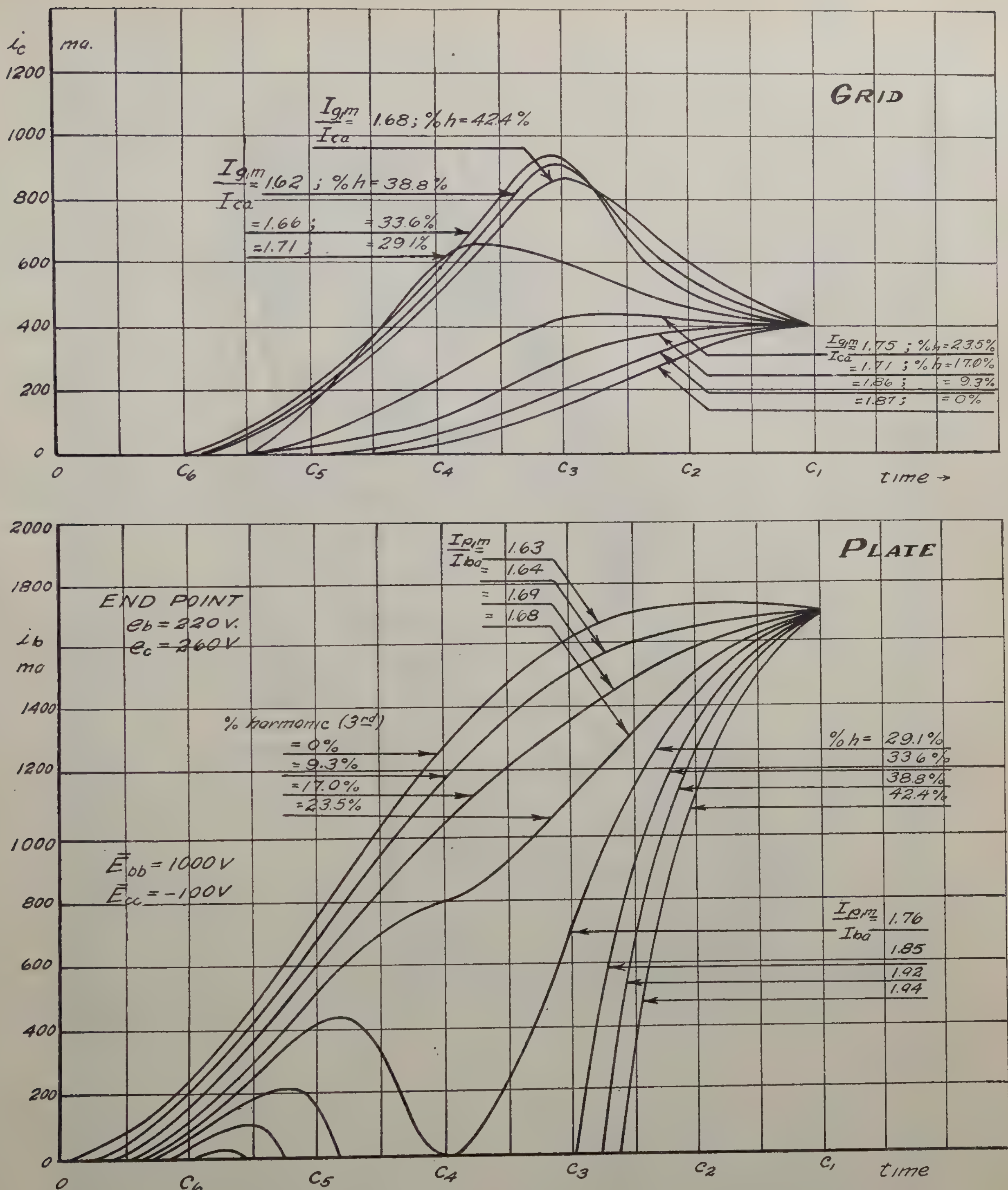
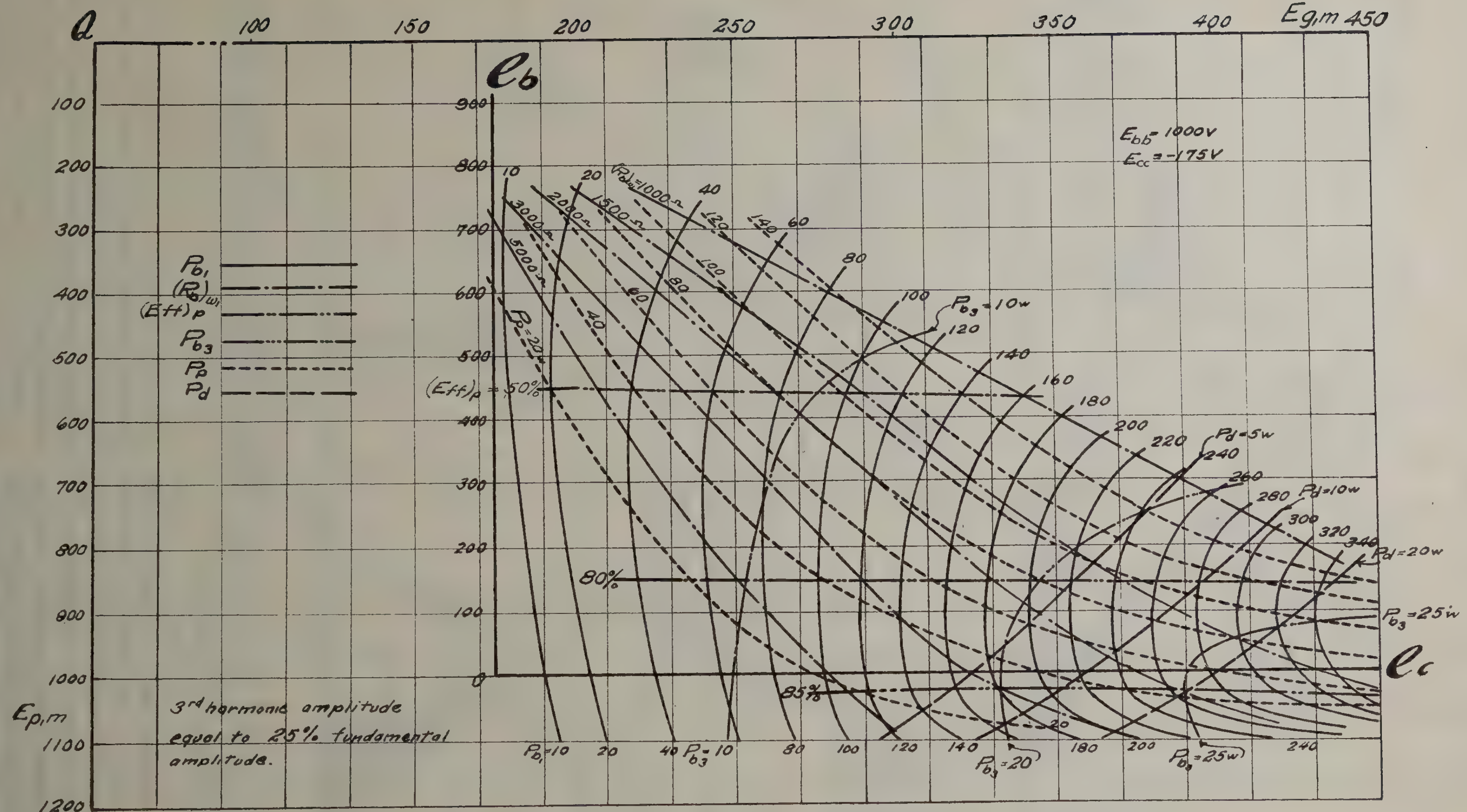


Fig. 9—Instantaneous-current pulses in the plate and grid circuits for various percentages of second-harmonic content in the fundamental plate voltage. The quiescent point Q is located at $E_{bb}=100$ volts, $E_{cc}=-100$ volts. The end point of the path of operation on the "power stroke" is located at $e_b=220$ volts, $e_c=260$ volts. Only half of the current pulses are shown. The pulses are symmetrical about the time $t=C_1$ on the diagram.

voltage used. These diagrams may be considered to be divided into two parts. To the right of the line $P_{b_3}=I_{p_3m}=0$ in the region designated B in these figures, the third-harmonic current is being delivered to the oscillator in the proper phase to supply the harmonic voltage correctly, and the path of operation that is obtained is shown in Fig. 13. To the left of the line $I_{p_3m}=0$, in the region designated A in these figures, the harmonic current is in the wrong phase to supply the harmonic voltage correctly, and the path of operation is as shown in Fig. 14. When the current I_{p_3m} approaches 0, the oscillator becomes unstable, and the path of operation jumps from the favorable to the unfavorable case, as shown in Fig. 15. When the path of operation is as shown in Fig. 14 the angle during which current flows is increased to almost 180 degrees, and P_p approaches



Dynamic Characteristics

Fig. 10—The dynamic characteristics of a 211 type tube when the third-harmonic voltage is introduced, in the correct phase, in the plate circuit. P_{b_1} is the fundamental power delivered to the plate tank circuit, $(R_b)_\omega$ is the equivalent resistance of the plate tank circuit, P_{b_3} is the third-harmonic power required to develop the necessary harmonic voltage, P_p is the plate dissipation, and P_d is the driving power. The third-harmonic voltage amplitude is equal in all cases to 25 per cent of the fundamental plate voltage.

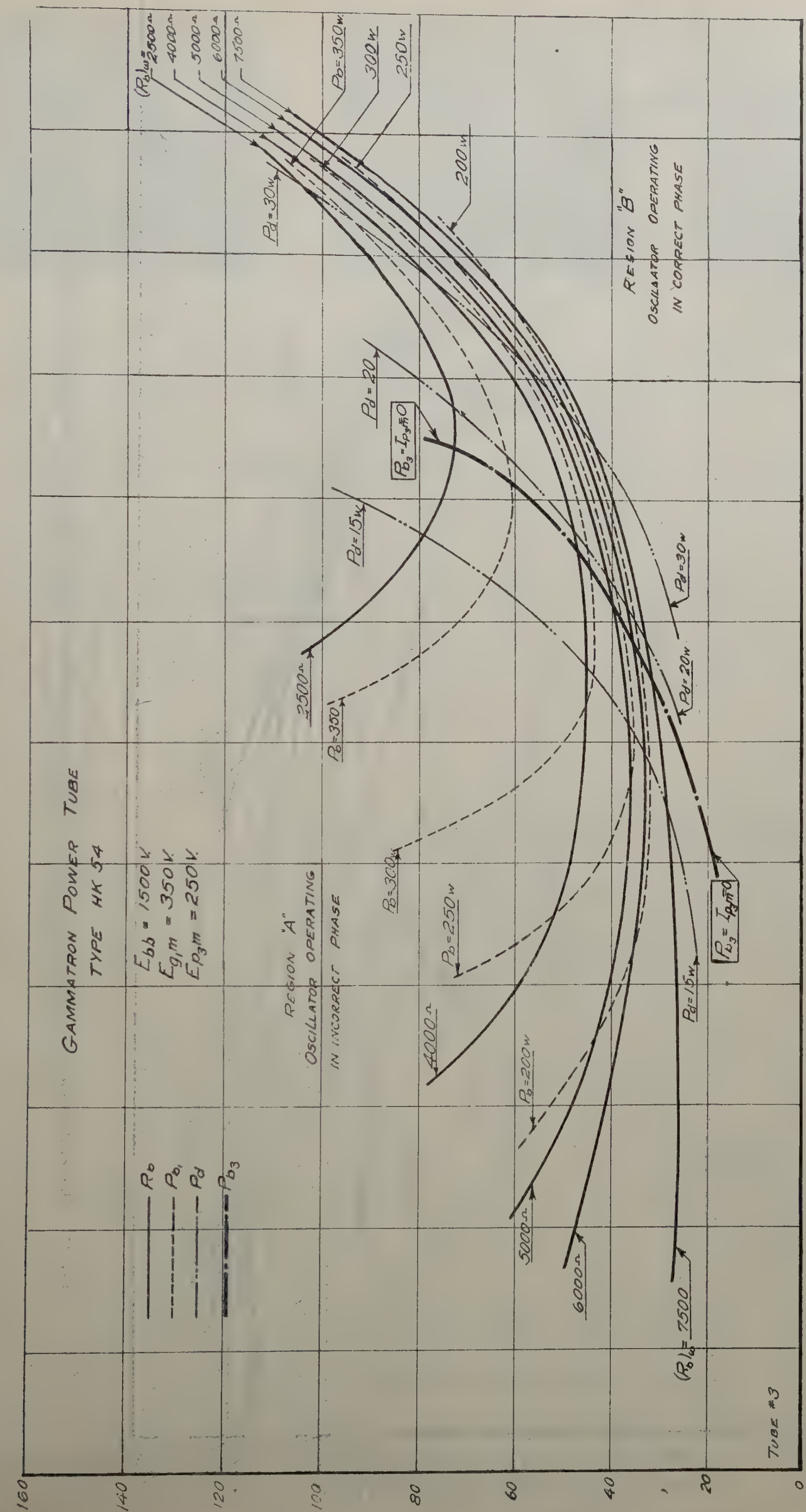


Fig. 11—Dynamic characteristics of a power amplifier equipped with a small auxiliary oscillator which supplies harmonic voltage to the amplifier. The oscillator frequency automatically locks in with the frequency of the input voltage to the amplifier. The oscillator voltage is in the correct phase in region *B*, incorrect phase in region *A*.

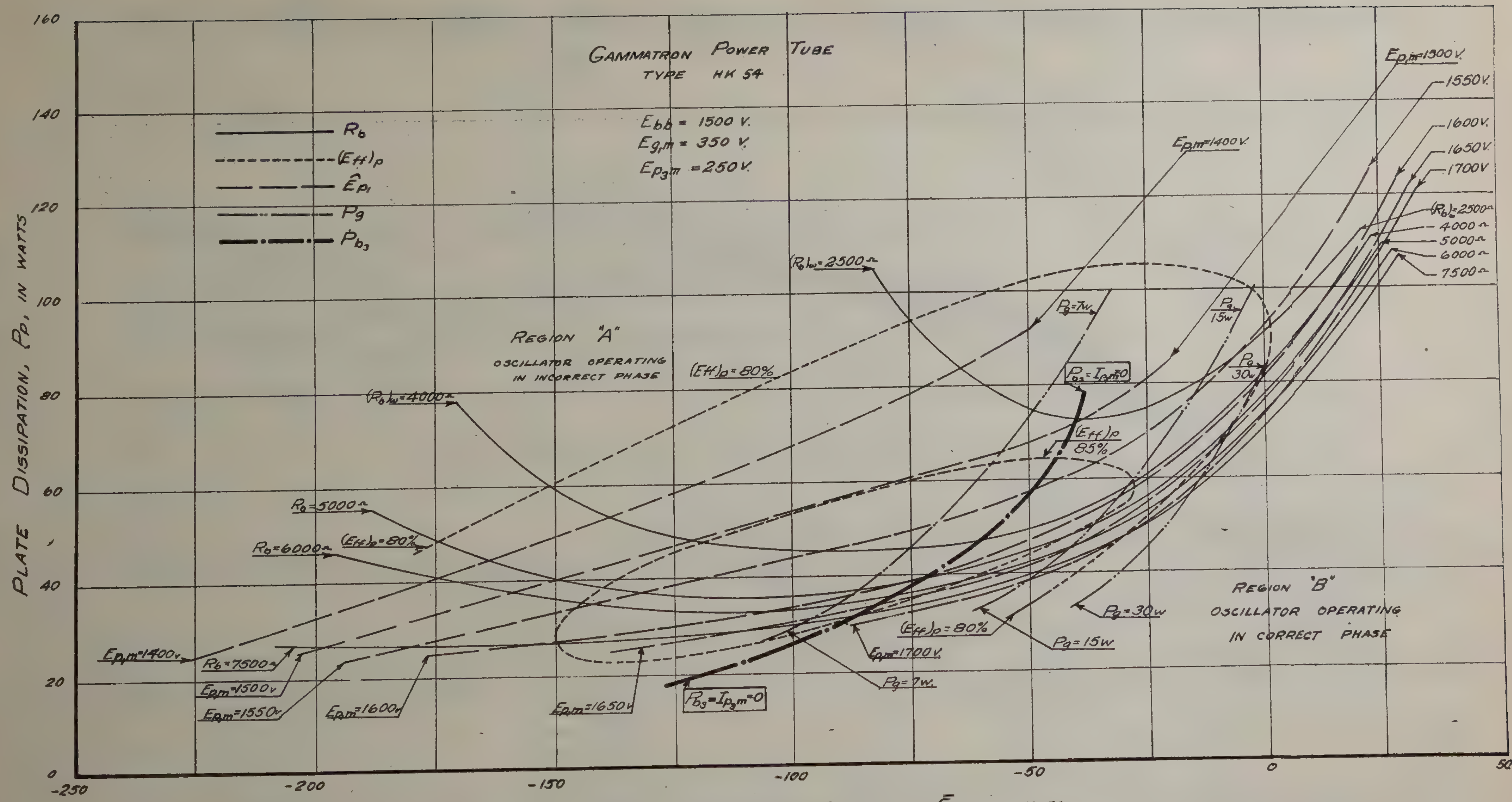


Fig. 12—Additional characteristics associated with those of Fig. 11. Lines of constant plate "equivalent" resistance have been reproduced from Fig. 11 for convenience. Regions A and B are separated by the line $P_{b_3} = I_{p_3m} = 0$.

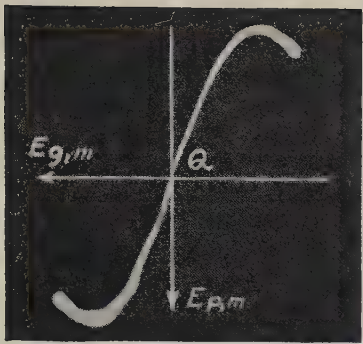


Fig. 13—Amplifier operating in region B, oscillator output in correct phase.

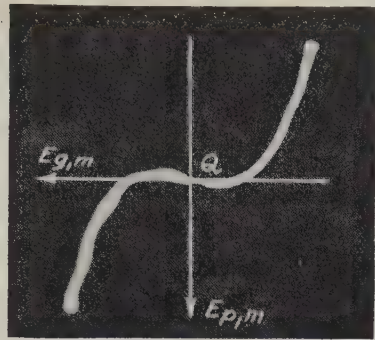


Fig. 14—Amplifier operating in region A, oscillator output in incorrect phase.

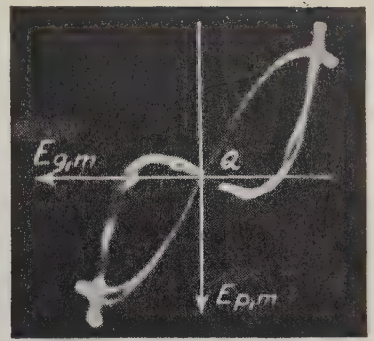


Fig. 15—Amplifier operating on line $P_{b_3} = I_{p_3m} = 0$ of Figs. 11 and 12. Oscillator unstable.

Figs. 13, 14, and 15—Oscillations of the path of operation taken for the amplifier whose characteristics are given in Figs. 11 and 12. The vertical deflecting plates of the oscilloscope were connected, through appropriate attenuators, between plate and filament of the amplifier tube. The horizontal plates were connected between grid and filament.

an extremely high value which may seriously injure the tube.

The ordinate of the diagrams of Figs. 11 and 12 might well have been E_{p_1m} instead of P_p. However, on this diagram, the limits of operation seemed to take the best form. Fig. 16 which is merely an enlargement of Figs. 11 and 12, shows more clearly the factors which bound the region of operation of the amplifier. In this figure we see that the region of operation is bounded on the left by the reversal in phase of I_{p_3m}, on the right, by excessive grid dissipation, and above by excessive plate dissipation. Within this region the variation of all the associated variables is indicated by the arrow direction. The black circle in the region of operation represents the most desirable position of operation. It is located slightly to the right of the I_{p_3m} line so as to allow for slight variations in the grid-polarizing potential due to line-voltage fluctuations, while at the same time giving the minimum grid loss and maximum power output.

When the independent variables were adjusted to

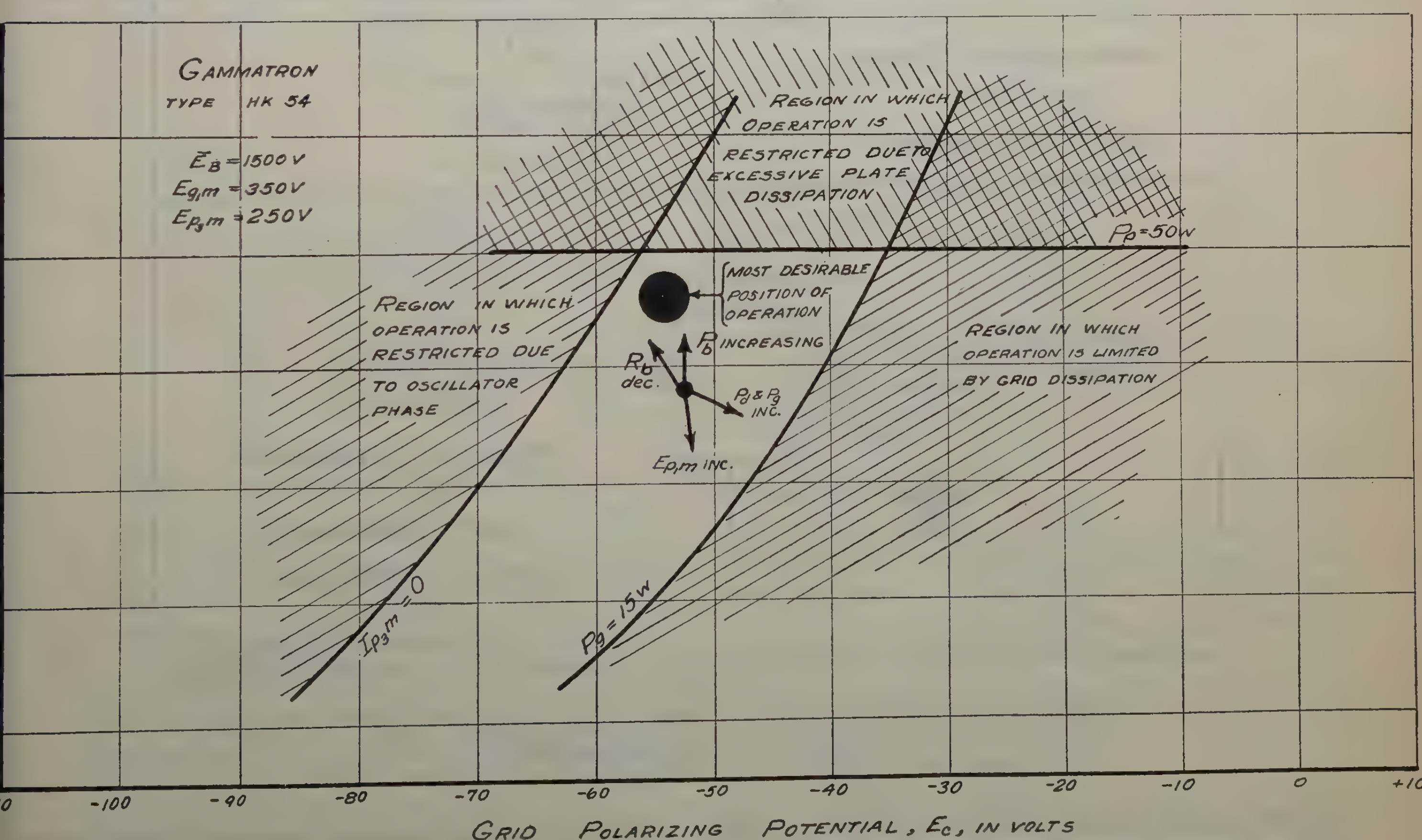


Fig. 16—Enlargement of the region in Figs. 11 and 12 where best operation of the amplifier takes place. The most desirable operating conditions of the system are indicated at the position of the black dot above.

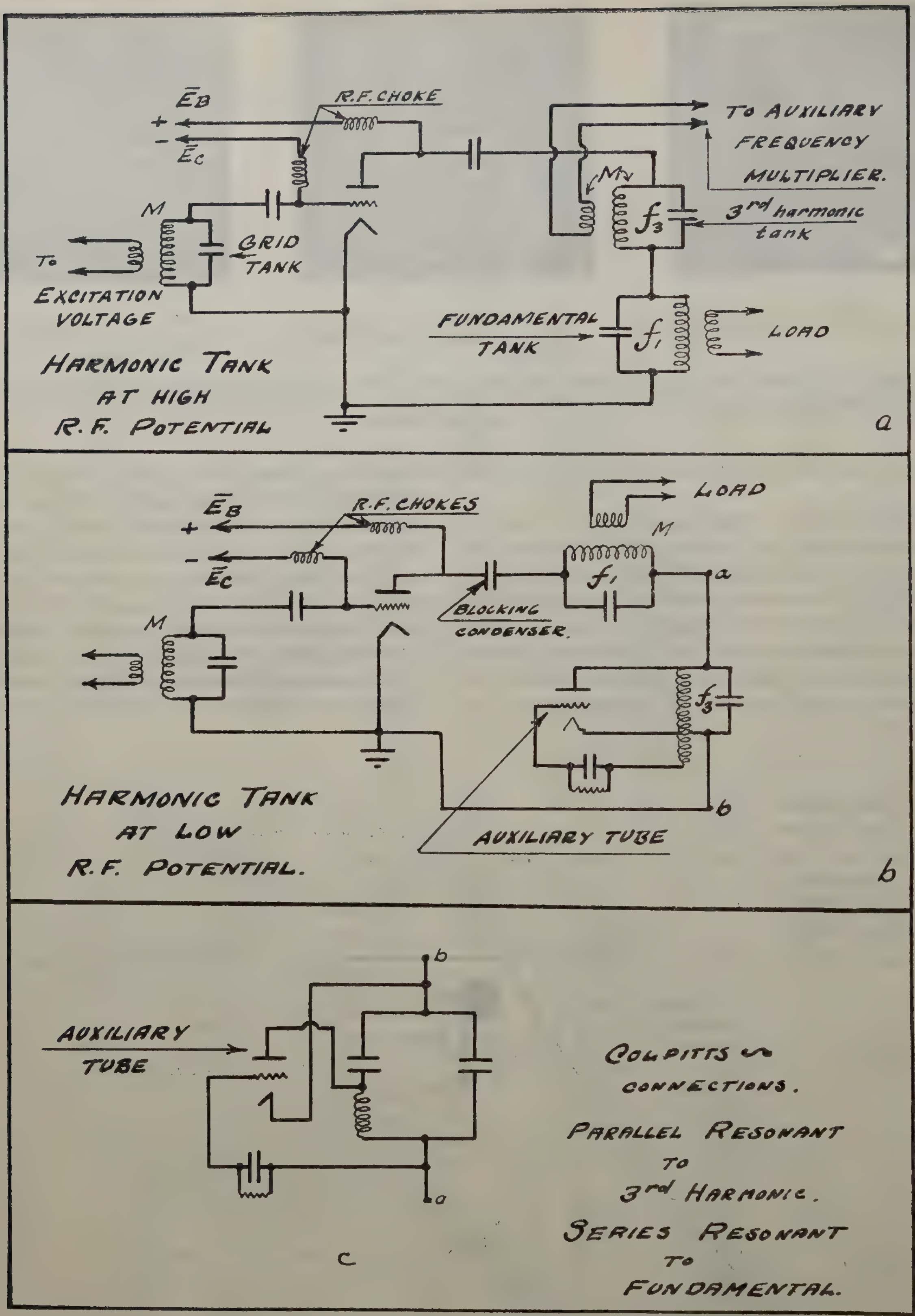


Fig. 17
(a)—Typical circuit for injection of harmonic voltage into the plate circuit of power amplifier.
(b)—An amplifier equipped with an auxiliary third-harmonic oscillator of the Hartley type for improvement of operating performance.
(c)—Arrangement of a Colpitts-type oscillator for use as an auxiliary oscillator.

give this position of operation with an oscillator supplying the harmonic voltage, the system was quite stable, and the power output of 310 watts at a plate efficiency of 87 per cent was obtained. When the direct-current power delivered to the oscillator was included, the efficiency was reduced to 85.5 per cent which shows the

small effect on the efficiency of the power consumed by the auxiliary oscillator. The optimum power output that could be obtained with the normal straight-line path for the same plate polarizing potential and driving power was 167 watts at 78 per cent efficiency.

Examples of circuit connections that may be used for high-efficiency amplifiers working on this principle are given in Figs. 17a and b. Fig. 17a shows the third-harmonic tank raised above ground by the radio-frequency potential of fundamental frequency across the plate-load tank circuit. In this case the oscillator must be link-coupled to the third-harmonic tank, as shown, since the filament supply of the tube in the auxiliary oscillator has high capacitance to ground. When the third-harmonic tank is lowered to ground potential, the tube of the auxiliary oscillator may be directly connected to it, as shown in Fig. 17b. In this case the plate tank is elevated above ground by the third-harmonic potential. Fig. 17c shows a Colpitts oscillator which may be installed between the terminals *a* and *b* of Fig. 17b in place of the Hartley oscillator. The tank circuit of the auxiliary oscillator is, in this case, so arranged as to bypass the fundamental current while at the same time it is parallel resonant to the third harmonic.

Fig. 18 shows a circuit which may be used to obtain the second- or third-harmonic voltage in the plate circuit in the correct phase. Fig. 19 shows a circuit arrangement which will generate the third harmonic in the correct phase with respect to the fundamental for efficient operation but will not work for the second harmonic. This is because a reversal in the fundamental voltage as fed to the auxiliary tube will reverse the phase of the third harmonic in the plate circuit of the tube but not the phase of the second harmonic. Fig. 20 shows a circuit which will work efficiently for the second harmonic but not the third. This is because, when the third harmonic is adjusted in phase to give the most favorable path for one of the "main" tubes, it will be in the wrong phase for the other. For the second harmonic, however, this is not the case and the "main" tubes are mutually benefited

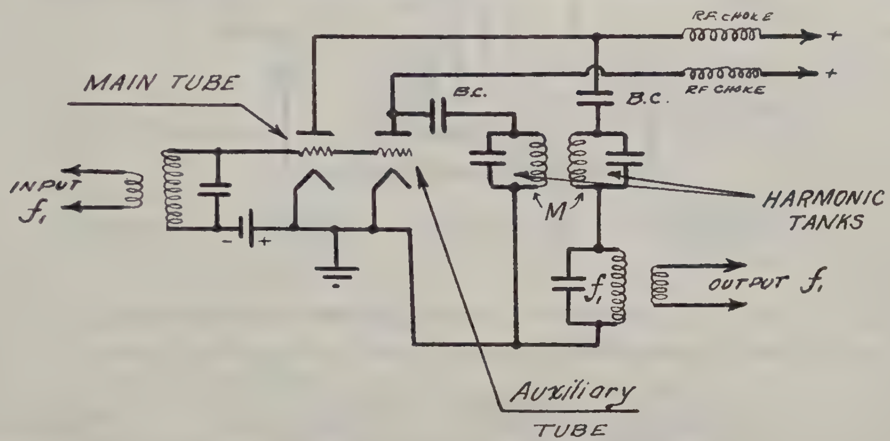


Fig. 18—Typical circuit for the generation of harmonic voltage in the correct phase to improve power-tube performance.

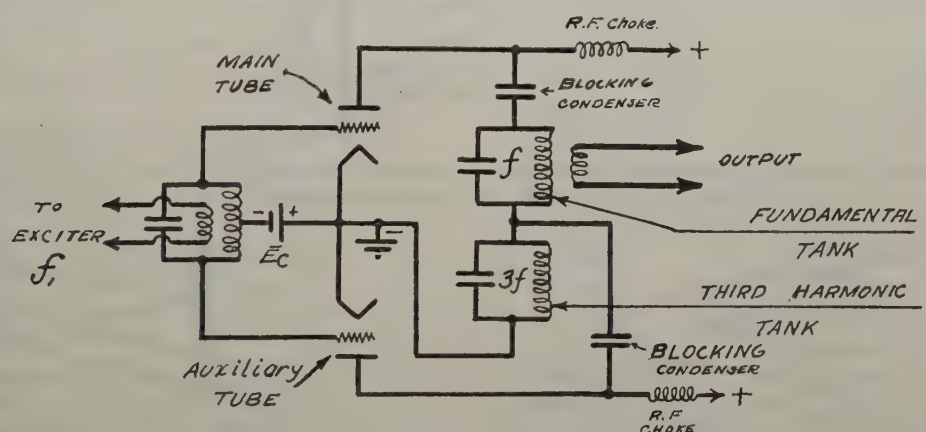


Fig. 19—This circuit will generate the third-harmonic voltage in the correct phase, but not the second-harmonic voltage.

In designing a frequency multiplier that would supply for this research any amount of harmonic power that might be desired, it was found that distinct improvement in frequency-multiplier design over the standard designs could be obtained.[7] The discussion of the most favorable path of operation for a tube is perfectly general. When we compare with this the path of operation for a frequency doubler or tripler, we find there is

[7] A comprehensive investigation of the influence of harmonic voltage on frequency multipliers has been carried out by J. E. Shepherd under the direction of E. L. Chaffee at the Cruft Laboratory, Harvard University. Results of this work will be presented at a later date.

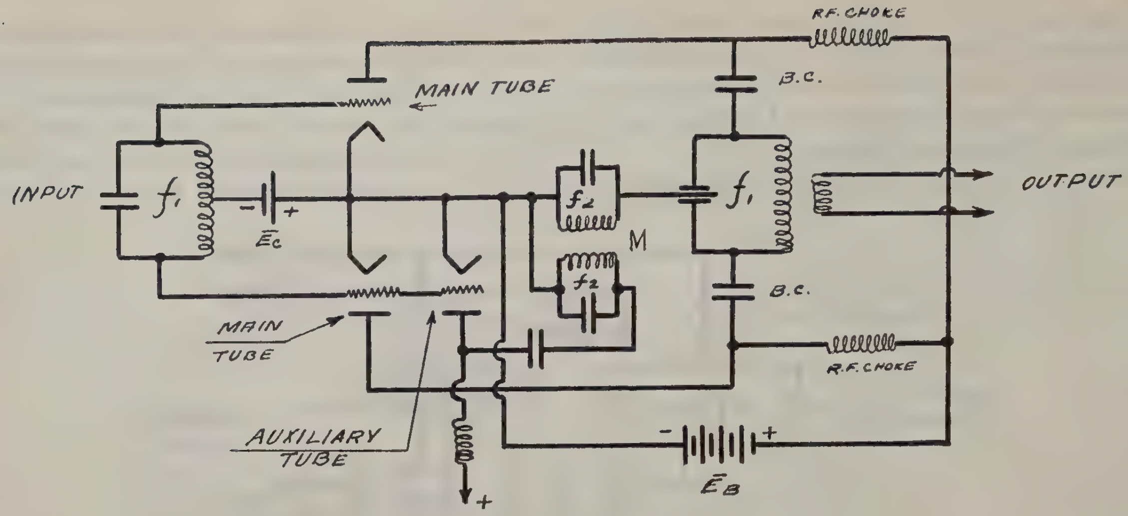


Fig. 20—Circuit suitable for the second harmonic but not the third.

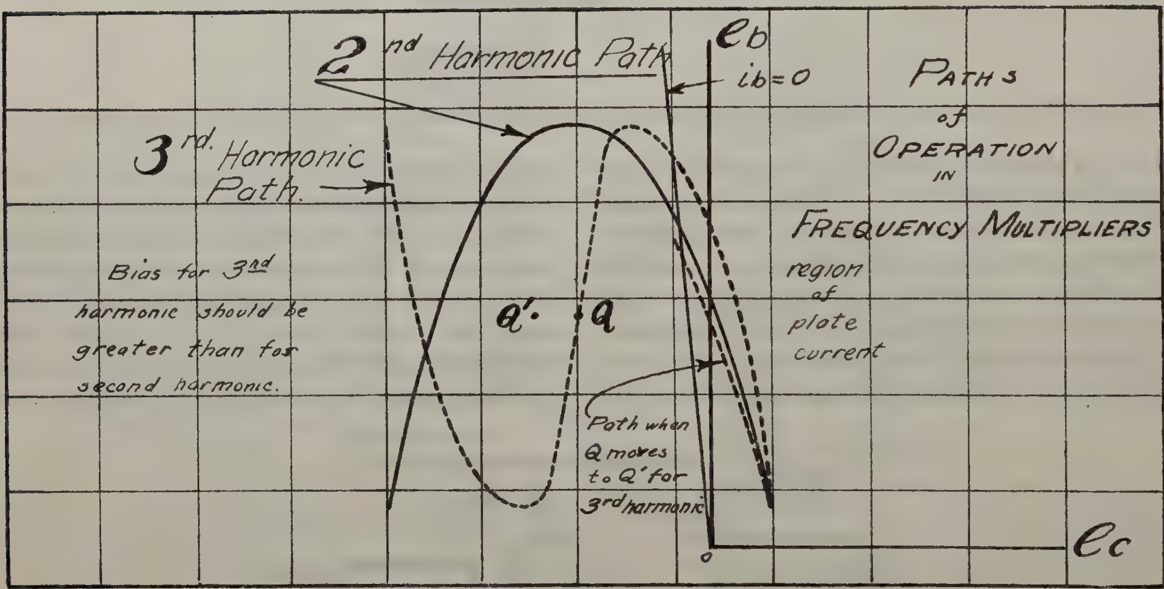


Fig. 21—Normal paths of operation in frequency doublers and triplers.

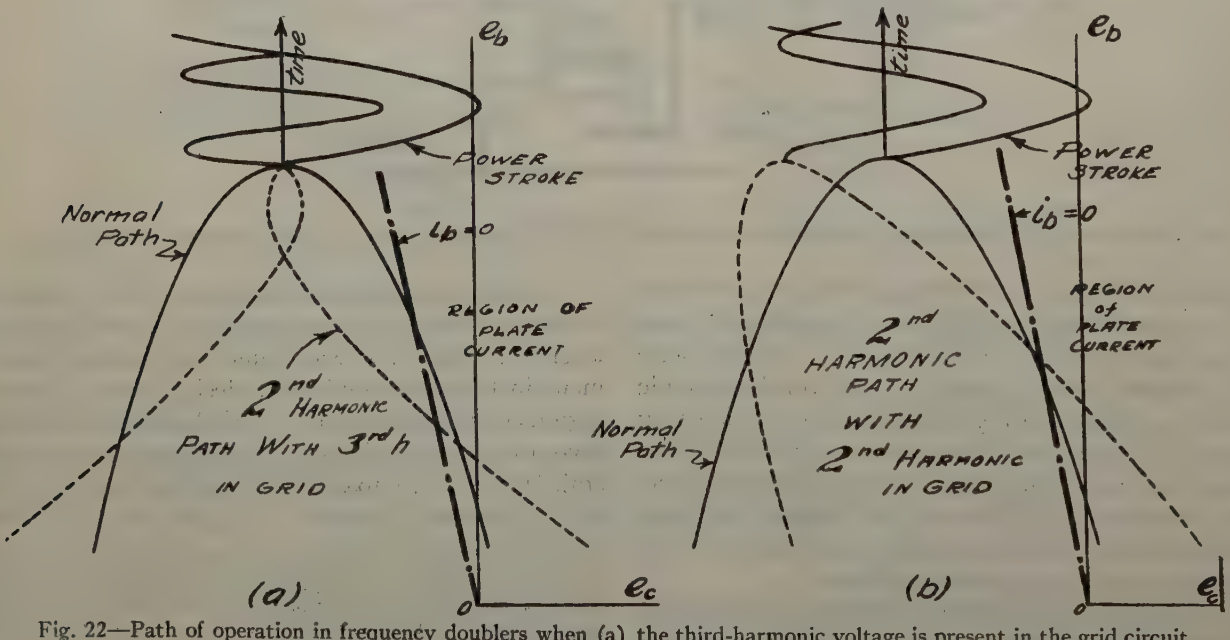


Fig. 22—Path of operation in frequency doublers when (a) the third-harmonic voltage is present in the grid circuit and (b), the second-harmonic voltage is present in the grid circuit.

opportunity for greater improvement than was found to be the case for the amplification of power of the same frequency. Fig. 21 shows paths of operation for a doubler or tripler. By the introduction of harmonic voltages in the grid circuit in the correct phase as shown in Fig. 22, it is possible to obtain very large gains in efficiency. This is evident, when we apply the arguments presented earlier, since the plate losses are reduced in even much greater proportion than in the case of the amplifier. By supplying the third-harmonic frequency to the grid circuit with a small auxiliary tank, the efficiency of the multiplier system was improved so as to approach the efficiency obtainable by ordinary class C amplification of power of the same frequency. Since very small amounts of second-harmonic power were required to accomplish this, the over-all gain is greatly increased.

By the correct adjustments of the operating angles of the exciter, which in this research was a crystal oscillator, it was possible to obtain in the plate circuit of this apparatus greater plate efficiency and at the same time a third-harmonic voltage for the frequency doubler without the use of an auxiliary tube. See Fig. 23. To obtain this operating angle, however, the power output of the exciter was slightly reduced. If an auxiliary tube is used to supply the third-harmonic voltage, the power

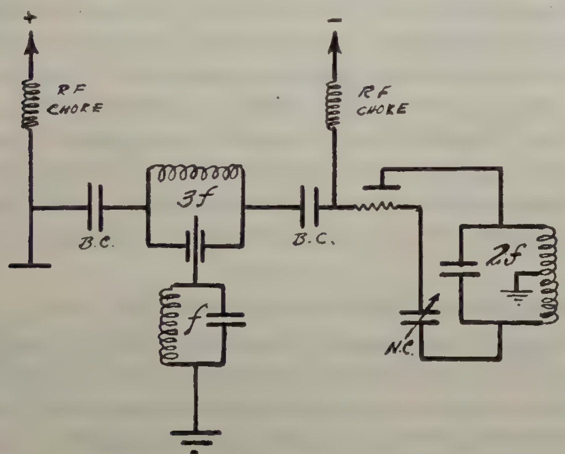


Fig. 23—Circuit for the generation of a harmonic voltage in the correct phase for a frequency doubler without the use of an auxiliary tube.

output of the system will be increased and also the third-harmonic voltage available for the doubler can be increased. See Fig. 24.

Since the feedback through the grid-plate capacitance of the frequency doubler is degenerative, a neutralizing condenser, as shown in Fig. 23, should be used. It is possible to make the doubler slightly regenerative and gain thereby, but since the effect of the neutralizing condenser is to shunt the grid tank circuit or circuits to ground, the aid due to regeneration is limited. It is possible that difficulties may arise due to self-oscillation of the doubler, when the feedback is too large.

Conclusion

It was found in this research that the most favorable path of operation for a vacuum tube is different from that which is obtained in the normal operation of class C amplifiers. It was found that a good approximation to this most favorable path of operation could be obtained by the introduction of the second- or third-harmonic voltages in the proper phase into the plate circuit of the

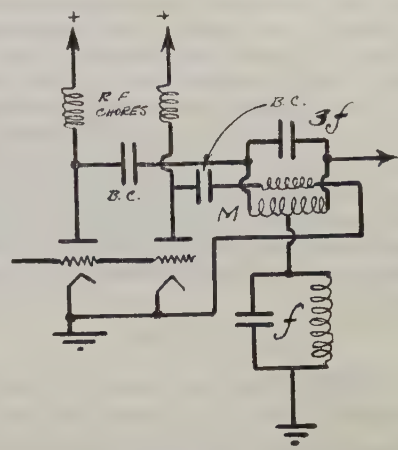


Fig. 24—Use of an auxiliary tube to improve the performance of a frequency doubler.

amplifying system. When these harmonics were introduced into the grid circuit, the path of operation was only slightly improved. Fourth- and higher-order harmonics were at best of little value in improving the path of operation and were detrimental in most cases, when introduced into either the plate or grid circuit.

The insertion of parallel resonant circuits tuned to the second- or third-harmonic frequencies into the grid circuit of amplifiers of this type, develop harmonic voltages which are in the wrong phase to improve the performance of the tube. The harmonic power necessary to develop these voltages must, therefore, be supplied by auxiliary equipment. These remarks are applicable to the plate circuit except in one special case, which has been pointed out in the text. In general, care must be taken to avoid excessively high instantaneous space currents, in order that reasonable filament life is not impaired.

In frequency doublers the third-harmonic voltage introduced into the grid circuit in the proper phase raises the efficiency of these units up to that obtained in ordinary class C amplification, even when the power required to supply the harmonic voltage is included. Similarly, high efficiency is obtained in frequency triplers.

A discussion of the effect of harmonic voltages in plate- and grid-modulated amplifiers and in linear amplification will be presented in a later paper.

Acknowledgment

It is a pleasure to have this opportunity to express my gratitude to Professor E. L. Chaffee for pointing out this problem and for his whole-hearted co-operation and guidance throughout its progress.

Information from his personal research has always been entirely at my disposal, and I have drawn from it freely. Much of this information has not as yet been published, and I wish to express appreciation for its use.

Coupled Antennas and Transmission Lines*

RONOLD KING†, ASSOCIATE, I.R.E.

Summary—The problem of coupled antennas is reviewed; unbalanced currents in and radiation from transmission lines are discussed briefly and resolved into transmission line and antenna problems. The use of sections of transmission line as coupling links between antennas is discussed qualitatively and illustrated in terms of a colinear array. The principles of phase reversing and of detuning stubs and sleeves are explained in terms of coupled circuits in which a section of transmission line may carry transmission-line currents, antenna currents, or both. Methods of driving transmission lines and of feeding antennas using transmission lines are discussed from the point of view of minimizing unbalanced currents on the line.

Coupled Antennas

THE TERM driven as applied to an antenna is used to designate an antenna that has two input terminals which are a negligible fraction of a wavelength apart and across which a potential difference is maintained by a generator connected either directly or through a transmission line. It is always implied even though not specifically stated in defining the input self-impedance of such an antenna at its terminals[1] that every other antenna (or any conducting or dielectric material) be sufficiently far from the driven antenna so that any change in its position which does not bring it nearer, in particular moving it very much farther away, produces no measurable difference in either the distribution of current[2] or the self-impedance of the driven antenna. Under these conditions the impedance at the terminals is correctly called the input self-impedance of the driven antenna, and every other conductor is individually only loosely coupled to it. It is important to bear in mind that this self-impedance includes as a major part the load due to radiation coupling of the driven antenna to currents which are ultimately made to flow elsewhere in the universe. These latter cannot, of course, be assumed absent in defining the self-impedance, but they are required to be far away.

If neighboring antennas are present, the input impedance of the driven antenna in question is no longer the self-impedance. The general mathematical problem of coupled antennas has been set up, but thus far only solved completely for the special case of a distant receiving antenna.[3] From the general integrals one can, however, conclude that the distribution of current along a driven antenna in the presence of a coupled antenna must differ from the distribution when it is very far from all other antennas. Accordingly, the input impedance of the antenna in question (antenna 1) when near a second antenna will differ from its value when isolated for two reasons. The first and most important is due to the potential difference appearing across the terminals of antenna 1 due to the current in antenna 2. The second is a change in the self-impedance due to the modified distribution of current. For present purposes it is sufficient to write the following equations (1) for the isolated driven antenna with input self-impedance Z_{00}, and (2) for the driven antenna in the presence of a second antenna as shown schematically in Fig. 1. In the latter case the self-impedances of the two antennas in each other's presence are, respectively, Z_{s1} and Z_{s2}; the mutual impedances are Z_{12} and Z_{21}

$$V_1^e = I_{01}(Z_1 + Z_{00}) \tag{1}$$

$$V_1^e = I_{01}(Z_1 + Z_{s1}) + I_{02}Z_{12} \tag{2a}$$

$$V_2^e = I_{01}Z_{21} + I_{02}(Z_2 + Z_{s2}). \tag{2b}$$

The superscript e on V designates an externally applied potential difference; the subscript 0 on the currents means the current at the input terminals. These are at the origin of the z axis which is oriented along the antenna. Z_1 and Z_2 are lumped impedances in series with antennas 1 and 2, respectively. In all cases not involving media of variable permeability and dielectric constant the following reciprocal relation is true:

$$Z_{12} = Z_{21}. \tag{3}$$

The equations (2) are exactly the same in form as the equations for two coupled circuits in ordinary network theory. By use of the principle of superposition the currents produced by each applied voltage may be determined separately, and the results combined algebraically. Accordingly, there is no loss in generality if V_2^e is set equal to zero. If V_2^e is actually zero, antenna 2 is said to be parasitic. If it is not actually zero its effect can always be calculated by interchanging the subscripts 1 and 2 and combining the solutions so obtained with those determined below. Thus the equations for one driven and one parasitic antenna are

$$V_1^e = I_{01}(Z_1 + Z_{s1}) + I_{02}Z_{12} \tag{4a}$$

$$0 = I_{01}Z_{21} + I_{02}(Z_2 + Z_{s2}). \tag{4b}$$

If a transmission line is connected between the input terminals AB of, say, antenna 1 and the output terminals of a conventional network containing a generator, as shown in Fig. 2, the symbols in (4a) must be changed as follows: One must write V_{AB} (open) for V_1^e with V_{AB} (open) meaning the open-circuit voltage across AB when the antenna is disconnected. Instead of Z_1 one must insert Z_{AB} where Z_{AB} is the impedance looking to the left (i.e., into the line) at AB with the

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† Cruft Laboratory and the Research Laboratories of Physics, Harvard University, Cambridge, Massachusetts.

[1] Ronold King and G. H. Blake, Jr., "The self-impedance of a symmetrical antenna," Proc. I.R.E., vol. 30, pp. 330–349; July, 1942.

[2] Ronold King and Charles W. Harrison, Jr., "The distribution of current along a symmetrical center-driven antenna," Proc. I.R.E., vol. 31, pp. 548–567; October, 1943.

[3] Ronold King and Charles W. Harrison, Jr., "The receiving antenna," Proc. I.R.E., to be published.

generator replaced by its internal impedance. It follows from Thévenin's theorem that (4a) is then a true equation.

The solutions of (4) may be derived directly from conventional analysis of coupled circuits. For present purposes it is sufficient to write down the expression for the impedance Z_{FG} (Fig. 1) offered to the generator. It is

$$Z_{FG} = (Z_1 + Z_{s1}) - \frac{Z_{12}Z_{21}}{(Z_2 + Z_{s2})} = Z_{11} - \frac{Z_{12}Z_{21}}{Z_{22}}. \qquad (5)$$

If one makes use of (3) and introduces the following notation:

$$Z_{12} = |Z_{12}|\, e^{j\theta_{12}}; \qquad \theta_{12} = \tan^{-1}\frac{X_{12}}{R_{12}} \qquad (6)$$

$$Z_{22} = |Z_{22}|\, e^{j\theta_{22}}; \qquad \theta_{22} = \tan^{-1}\frac{X_{22}}{R_{22}} \qquad (7)$$

one readily obtains

$$R_{FG} = R_{11} - \frac{|Z_{12}|^2}{|Z_{22}|}\cos(2\theta_{12} - \theta_{22}) \qquad (8a)$$

$$X_{FG} = X_{11} - \frac{|Z_{12}|^2}{|Z_{22}|}\sin(2\theta_{12} - \theta_{22}). \qquad (8b)$$

It follows directly that the input impedance Z_{AB} of

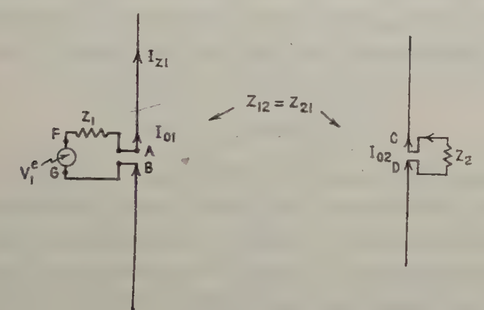


Fig. 1—Schematic circuit for coupled antennas.

antenna 1 in the presence of antenna 2 has the following resistive and reactive parts:

$$R_{AB} = R_{s1} - \frac{|Z_{12}|^2}{|Z_{22}|}\cos(2\theta_{12} - \theta_{22}) \qquad (9a)$$

$$X_{AB} = X_{s1} - \frac{|Z_{12}|^2}{|Z_{22}|}\sin(2\theta_{12} - \theta_{22}). \qquad (9b)$$

These relations reduce to those of transformer coupling in ordinary circuits if R_{s1} and X_{s1} apply to the primary coil if one writes $Z_{12} = j\omega M$ so that $\theta_{12} = \pi/2$. In this case one has for the transformer

$$R_{AB}(\text{trans}) = R_{s1} + \frac{\omega^2 M^2}{|Z_{22}|}\cos\theta_{22} \qquad (10a)$$

$$X_{AB}(\text{trans}) = X_{s1} - \frac{\omega^2 M^2}{|Z_{22}|}\sin\theta_{22}. \qquad (10b)$$

An important difference exists between the general case of two circuits coupled by the mutual impedance between antennas and two inductively coupled circuits of conventional type. Whereas R_{in}(trans) as given by (10a) is always larger than R_{s1} by the resistance "reflected" into the primary from the coupled secondary, R_{AB} for the antenna as given in (9a) may be larger or smaller than R_{s1} depending on the distance between the antennas. In the case of the transformer the coupled secondary always represents an increased load on the primary due to the power which is dissipated in the secondary and which must be supplied by the primary. With the antenna the presence of a neighboring antenna necessarily implies an increase in the load corresponding to the very small amount of power dissipated in heat in the coupled antenna, but it also means a change in radiated power. This is manifested by a complete or partial cancellation in certain directions of the fields due to the currents in the two coupled antennas. This may be compensated by greater or smaller increases in other directions, so

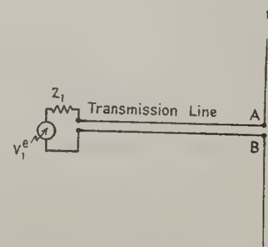


Fig. 2—A symmetrical antenna driven from a parallel-wire transmission line.

that the input resistance R_{AB} may be either larger or smaller than the self-resistance R_{s1} depending upon whether the presence of the closely coupled antenna increases or decreases the total power radiated. Analytically this is determined by the algebraic sign of $\cos(2\theta_{12} - \theta_{22})$. It is well to note that an antenna which is self-resonant, i.e., adjusted to resonance when no other antenna is near, is no longer resonant if another antenna is brought near and the mutual impedance between the two is not negligibly small. The same conclusions are true for more conventional coupled circuits.

Mutual Impedance

The mutual impedance (referred to input current) of one antenna in the presence of another depends upon the size, shape, and relative orientation of the conductors forming the two antennas as well as upon their conductivities. The analytical problem of deriving formulas for the mutual impedance of antennas has not up to the present time been solved either rigorously or to a reasonable degree of approximation even for the simplest case of two identical, parallel, center-fed antennas of small radius. The data for mutual impedance which are available have all been calculated on the assumptions, first, that the distribution of current along a driven antenna, (and even of a parasitic antenna), when in the presence of another is just the same as that of an isolated driven antenna and second, that the simple sinusoidal distribution of current (which is correct only along an infinitely thin, perfectly conducting antenna) is a good approximation. Except when[2] $X_1 + X_{s1} = 0$, this assumption may be so very far

from correct that the results which are based upon it are often not even rough approximations. A more accurate analysis depends upon the determination of the distributions of current and charge along antennas coupled in different ways. This is no simple problem. For reasons which have been discussed[2] the mutual impedance between two parallel and identical center-driven antennas of half-length $h=\lambda/4$ (or odd multiples thereof), and extremely small radius may be obtained approximately from calculations based on the simple sinusoidal distribution of current. Curves for $|Z_{12}|$ and θ_{12} for this special case are in the literature.[4] As the separation b between the antennas is made to vanish, Z_{12} approaches, as it should, the expression for self-impedance calculated under the same assumptions. It is

$$Z_{s1} = (Z_{12})_{b=0} = 73.13 + j42.5. \tag{11}$$

On the other hand for a very thin antenna ($a/\lambda = 10^{-5}$) with $h=\lambda/4$ one has[1]

$$Z_{s_1} = 69.5 + j36; \tag{12}$$

for a moderately thick antenna ($a/\lambda = 10^{-3}$) with $h=\lambda/4$ one has

$$Z_{s1} = 66.5 + j31. \tag{13}$$

While the discrepancies between (11) and the practically possible values (12) and (13) (as well as those for all intermediate thicknesses) is considerable, a rough estimate of the mutual impedance of extremely thin antennas is presumably given by assuming a sinusoidal current. It is important to note, however, that this is only true for $h=\lambda/4$ *and not at all for self-resonant antennas* however thin, of length h_r less than $\lambda/4$. For a resonant antenna X_{s1} is zero and this cannot be approximated by 42.5. Until satisfactory data for mutual impedances and for self-impedances of antennas in the presence of others are made available, precise calculations of impedances in coupled antennas of this simple type are impossible. The accurate calculation of self- and mutual impedances in simple parallel arrays is one of the major problems awaiting solution in the theory of antenna circuits. Once these are known many important circuit problems relating to coupled antennas can be solved numerically as well as formally. Only qualitative discussions of such problems are now possible, except for extremely thin identical antennas with $h=\lambda/4$, or with $X_1+X_{s1}=0$.

The Coefficient of Coupling between Antennas

The input impedance of an antenna (1) in the presence of a single antenna (2) is given by

$$Z_{AB} = Z_{s1} - \frac{Z_{12}Z_{21}}{Z_{22}}. \tag{14}$$

This may be rearranged as follows:

$$Z_{AB} = Z_{s1}\left(1 - \frac{Z_{12}Z_{21}}{Z_{s1}Z_{s2}}\cdot\frac{Z_{s2}}{Z_{22}}\right). \tag{15}$$

[4] G. H. Brown and Ronold King, "High-frequency models in antenna investigations," Proc. I.R.E., vol. 22, pp. 457–480; April, 1934; curves marked $G=90°$ in Fig. 13.

Let a complex coefficient of coupling between antennas 1 and 2 be defined by

$$k^2 = \frac{Z_{12}Z_{21}}{Z_{s1}Z_{s2}}. \tag{16}$$

Then

$$Z_{AB} = Z_{s1}\left(1 - k^2\frac{Z_{s2}}{Z_{22}}\right). \tag{17}$$

Loose coupling may be defined by

$$|k|^2 \ll 1. \tag{18}$$

This condition is always satisfied by antennas which are sufficiently far from each other. The coefficient of coupling approaches but does not quite reach unity if the two antennas are self-resonant and are moved very close together. The resulting circuit closely resembles the parallel line. It is clear that all degrees of coupling are possible with antennas, and that effects may be anticipated similar to those for coupled circuits of more conventional types. In particular, double-resonance peaks for greater than critical coupling may be expected. Two important differences obtain. These are first, the fact that mutual resistances and reactances may be negative as well as positive, and second, that resonance peaks are always blunt due to radiation, unless the antennas are very close together. Extensive and accurate quantitative results of coupled-circuit effects in antennas are not yet available either from theoretical or experimental investigations.

Transmission Lines

Conventional analysis of two-conductor transmission lines, whether of the parallel-wire or coaxial type, assumes that the currents in the two conductors are equal and opposite. The solutions obtained are valid only if this condition is actually fulfilled. Although the ideal condition of operation for transmission lines is precisely that assumed in the analysis, it does not follow that it must necessarily obtain in practice unless special precautions are taken. Currents in the two conductors of a transmission line will be equal and opposite only if the driving generator, the load, and the line itself are symmetrically constructed and connected. The requisite type of symmetry is different for parallel-wire and coaxial lines, and the consequences of asymmetry are not alike though equally undesirable from the point of view of transmission of power. They will be described and discussed below.

Parallel wires are not always used as in two-wire transmission lines. Thus in a cage antenna several conductors are arranged in parallel with approximately equal currents in the *same direction*. In a four-wire line currents are in the same direction in one diagonal pair, in the opposite direction in the other. If a parallel-wire line is driven and loaded symmetrically but located parallel to the electric field of an antenna, it will carry equal and opposite currents $I_{1t} = -I_{2t}$ due to the generator, and in addition practically equal currents in the same direction $I_{1a} = I_{2a}$ due to the electric field. The former may be called *transmission-line currents*, the

latter, *antenna currents*. The line may be adjusted to carry resonant transmission-line currents I_t by suitably placing a conducting bridge; it may be arranged to carry resonant antenna currents I_a by adjusting its over-all length. A line which carries both transmission-line and antenna currents has a complicated distribution of current given by

$$I_1 = I_{1t} + I_{1a} \tag{19a}$$
$$I_2 = I_{2t} + I_{2a} = -I_{1t} + I_{1a}. \tag{19b}$$

Such a distribution is said to be unbalanced. Every unbalanced distribution of current in a parallel-wire line may be expressed in terms of transmission-line and antenna currents. The former may be analyzed using line theory; the latter using antenna theory. The former contribute a negligible amount to radiation (if the line is closely spaced as it must be if conventional line theory is to be used); the latter radiate as for any other antenna. If a parallel line radiates appreciably it is because it carries an antenna current.

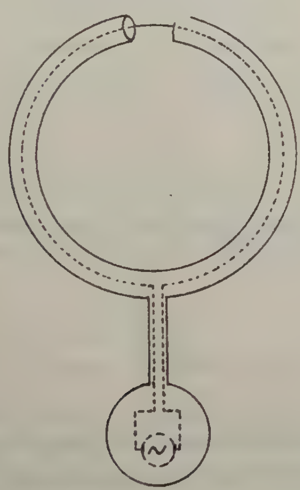


Fig. 3—Use of a coaxial line to form a shielded loop antenna.

Coaxial lines may also carry antenna currents. In fact, it is more difficult to avoid antenna currents on a coaxial line due to an unsymmetrical generator or load than on a parallel line as will become apparent below. In the case of a coaxial line the antenna currents flow on the outside surface of the outer conductor, the transmission-line currents on the inner surface of the outer and the outer surface of the inner conductor. On a transmission line such antenna currents are always undesirable because they radiate power just as currents on the surface of any other antenna. On the other hand in the shielded-loop antenna, which consists essentially of a coaxial line as shown in Fig. 3, the entire operation as an antenna depends on the antenna current which flows on the outer surface.

The Coupling of Antennas and Transmission Lines

In carrying out the mathematical analysis of two coupled antennas no assumption whatsoever was made regarding the relative orientations or positions of the coupled antennas. The diagrams of Fig. 1 suggested that they might be parallel, and this is actually often the case, as in so-called parallel arrays. But other orientations, as for example, antennas placed end to end along a single axis are also possible and useful. Instead of coupling antennas only through the mutual impedance between them, additional coupling especially in the form of sections of transmission line (so-called phase-reversing stubs) is often provided for special purposes. Furthermore antennas are not always driven by potential differences maintained between two symmetrically placed terminals. Frequently they are asymmetrically driven by being coupled more or less closely at one end to each other or to a resonant or anti-resonant section of a transmission line. The mathematical analysis of all such cases depends on the intricate problem of providing data on mutual impedances, not only of the antennas but also of the sections of transmission line. Consequently, calculation of the input impedance at the terminals of an antenna or of a transmission line, which is coupled in one way or another to one or more antennas, is not at present possible. In some cases the radiation resistance R_m^e referred to a maximum sinusoidally distributed current has been

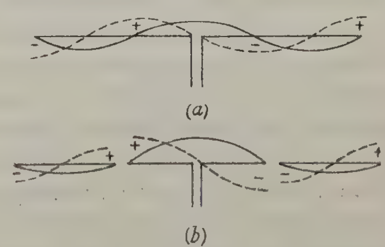


Fig. 4—(a) Center-driven antenna of half-length $h \doteq 3\lambda/4$.
(b) Center-driven dipole with coupled colinear elements. Solid curve: distribution of maximum current for infinitely thin antenna; dotted curve: distribution of charge a quarter period after the indicated current.

computed. For practically available antennas the actual distribution of current may depart sufficiently from the assumed sinusoidal one to lead to errors[2] in R_m^e as large as 50 per cent. A rough comparison of radiated power is, nevertheless, possible in terms of R_m^e.

In attempting to understand in a qualitative way the operation of, and the distribution of current in resonant systems consisting of several coupled components of which at least one is an antenna, it is well to bear in mind that one is usually concerned only with highly conducting antennas and lines in which the charges are free to redistribute themselves continuously in such a way that at every instant and at every point along each conducting surface the interacting tangential forces practically cancel. Accordingly, any redistribution of charge which follows the coupling of two circuits must be of such a kind as to lead to a virtual cancellation of tangential forces along all the conductors. In circuits consisting of several more or less closely coupled components there may be more than one possible mode of oscillation. The mode which

is actually excited depends upon the relative directions and magnitudes of the tangential forces due to the several interacting parts of the circuit and upon the power radiated in each case.

The Colinear Array as a Coupled Circuit; Phase-Reversing and Detuning Stubs

A number of important properties of coupled antennas may be described in terms of the colinear array. For simplicity let this consist of three identical self-resonant units each something less than a half wavelength long arranged end to end as shown in Fig. 4a. Only the central unit is driven from a resonant section of transmission line. A number of different methods of coupling the three units together are pictured in Figs. 4 and 5, and these will be discussed in turn. The approximate distributions of the principal components of the maximum current and the charge per unit length a quarter period later are shown respectively in solid and dashed lines. In Fig. 4a the three units are connected directly together, the coupling between them is so close that they form in effect a single resonant antenna of length near $3\lambda/2$.

In Fig. 4b the two outer antennas are shown pulled away from the central one, thus decreasing the coupling. To a first approximation the distributions of current and of charge per unit length along each of the three units is changed relatively little as the separations are increased from very small values. The amplitudes in the outer units decrease rapidly, however. Adjustments in the lengths of the outer antennas or of the central unit produce coupled-circuit effects on the amplitudes of current that resemble those encountered in conventional coupled circuits with lumped constants as the secondary or primary tuning is varied. Depending upon the degree of coupling as determined by the spacing, double or single resonance peaks may be obtained. The load on the generator driving the central unit is not correspondingly diminished when less power is transferred to the outer antennas as these are moved outward. As the degree of coupling of the central unit to the outer two is decreased less power is supplied to the two adjacent antennas, but more is radiated from the central unit. This is readily understood from the fact that the currents in the parasitic antennas are both opposite in direction to that in the central unit, so that a partial cancellation of the fields due to them in most, but not necessarily in all, directions must result. The radiation resistance referred to maximum sinusoidal current for the case of Fig. 4a is 105.5 ohms which is considerably less than three times the corresponding value, 73.1, for the central unit alone.

In Fig. 5a a quarter-wave bridged-end section of parallel line is connected to the adjacent terminals of the antennas shown in Fig. 4b. There are now five resonant circuits. The three antennas are coupled together just as before in Fig. 4b, but they are now also coupled to the upper ends of the stub sections of transmission line. Very little power would be required to drive only the stubs as coupled secondaries of the central antenna because they can radiate only a negligible amount of power as a result of the cancellation of forces due to their equal and opposite transmission-line currents. The currents will be equal and opposite because the over-all length of the stubs is not such as to permit significant antenna currents to be superimposed on the transmission-line currents. But when the outer antennas are attached, the situation is different. The only possible condition of resonance of all the coupled circuits in Fig. 5a requires equal and opposite currents and charges in the stubs, and this reverses the currents and charges in the outer antennas as compared with those in Fig. 4. Thus the coupling forces between the

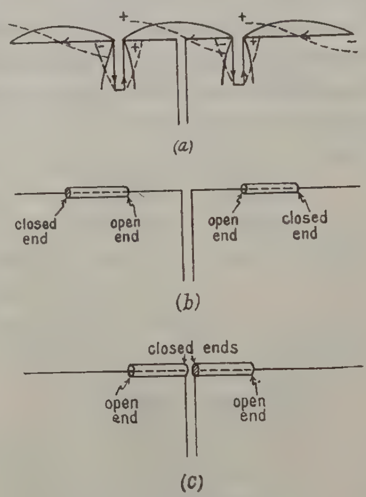
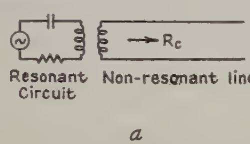


Fig. 5—Colinear arrays.

stubs and the outer antennas—forces which are confined to short distances near the points of contact—are in opposition to those existing directly between the antennas which seek to produce the distribution of current and charge of Fig. 4. But so long as the stubs are adjusted to resonance, the forces between them and the antennas are very much the stronger and the distribution shown in Fig. 6a prevails. It is to be noted that the currents in all three antennas are now in the same direction. This means that the field of the colinear array of three units is enormously increased because cancellation of forces due to currents in different parts of the array is very much reduced in more directions than those in which it is increased. Accordingly very much more power must be supplied to the colinear array of Fig. 5a than to the antenna of Fig. 4a or the coupled array of Fig. 4b for the same input current. That is, the input resistance at resonance is much greater for the colinear array than for the straight antenna of the same length. An estimate of the difference is obtained by comparing the radiation resistances referred to maximum sinusoidal current in the two cases. For Fig. 4a $R_m{}^e$ has been given to be 105.5 ohms; for Fig. 5a it is 316.5 ohms. As a radiating device the colinear array is thus three times as effective as the linear radiator of the same length, $3\lambda/2$, and over four

times as effective as the central unit alone. Arrangements equivalent to that of Fig. 5a are shown in Figs. 5b and 5c. Coaxial sleeves have been substituted for parallel-line stubs. Their operation is described in detail in the following section dealing with a completely coaxial colinear array.

It is important to bear in mind that the colinear arrangement of Fig. 5 does not behave like a single tuned circuit as does the antenna of Fig. 4a. It consists of five more or less closely coupled resonant circuits, and its behavior is correspondingly complex. If the tuned circuit analog (*but not equivalent*) of a single antenna is taken to be a series-resonant circuit coupled to an infinite line, as shown in Fig. 6a, then the corresponding analog of the colinear array approximates the circuit of Fig. 6b. The characteristic resistances R_c of the nonresonant lines of Fig. 6 are taken to be approximate



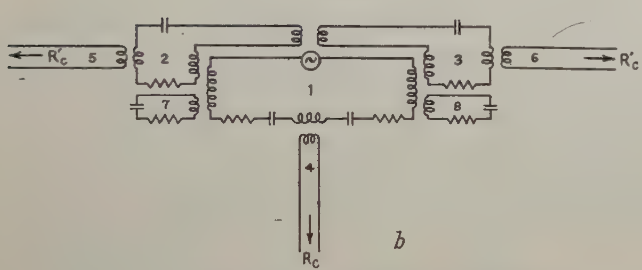


Fig. 6—(a) Circuit analog (not equivalent) of a single driven antenna.
(b) Circuit analog (not equivalent) of a three-element colinear array.
1. Analog of central antenna element of Fig. 5.
2. Analog of left antenna element of Fig. 5.
3. Analog of right antenna element of Fig. 5.
4. Analog of radiation load on central antenna element.
5. Analog of radiation load on left antenna element.
6. Analog of radiation load on right antenna element.
7. Analog of left phase-reversing stub of Fig. 5.
8. Analog of right phase-reversing stub of Fig. 5.

analogs of the characteristic resistance $R_c = 376.7$ ohms of space to which power may be assumed to be transferred in the radiation process. It is clear that complicated coupled-circuit effects must be expected. In particular, the colinear array cannot be tuned to resonance merely by adjusting the length of the two outer ends as can the antenna of Fig. 4a. Each of the five circuits must be tuned separately, and multiple resonance peaks of current due to greater than critical coupling may be expected unless the spacing of the parallel or coaxial conductors in the stubs is too small. Because of the large load due to radiation all resonance peaks are blunt.

If one of the outer antennas is detuned, the amplitude of current in it will, of course, decrease to a small value, but a slight readjustment in the tuning of the rest of the circuit will keep this in resonance with no appreciable change in the distribution of current (if the antennas are very thin). For the same input current the radiated power will diminish, as will the input resistance. For the same applied voltage the amplitudes of current in all but the detuned antenna will increase.

If the stubs are detuned sufficiently either by changing their length or by decreasing the spacing of the parallel or coaxial conductors the periodically varying high concentrations of charge at their upper ends will decrease in amplitude, and the voltage induced in the outer antennas by tangential forces along the surface of the attached ends will decrease. (The induced re-

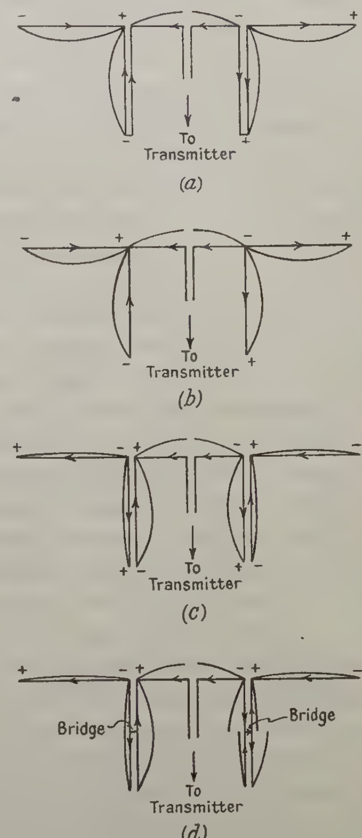


Fig. 7—Antennas and lines coupled in various ways.

distribution of charge cancels these.) If the voltage induced in the outer antennas as a result of their coupling to the attached stub is reduced below the opposing voltage induced as a result of their respective coupling to the central unit, then a condition similar to that of Fig. 4 is established. If the stubs are completely removed, or if their length is increased to nearly one-half wavelength as in Fig. 7a, resonance can be restored. In the latter case each stub *acts essentially as a single conductor*, with both wires in parallel carrying antenna currents in the same direction. There will be a maximum concentration of charge at the bottom a quarter period after maximum current halfway down. Transmission-line currents will be extremely small because of the low input impedance across which the driving potential difference would have to be established. In fact if the two conductors of the stub are

connected at the top or are actually replaced by a single conductor as in Fig. 7b, no significant change in the mode of oscillation or the distribution of current occurs. One now has the equivalent of five coupled antennas, three horizontal, two vertical, and a corresponding change in the radiation from the array. If each half-wave stub is a section of coaxial line, closed at the bottom by a metal disk connecting the inner and outer conductors, practically no transmission-line current will flow in the completely detuned interior. On the other hand the outer surface is simply a thick antenna which is self-resonant at a length somewhat shorter than $\lambda/2$.

If the terminating bridge (or disk) at the bottom of the stub is removed, the parallel line (or the two-conductor interior of the coaxial line) has a high input impedance at a point of maximum concentration of charge on the antenna, so that equal and opposite currents on the two conductors with the distribution of Fig. 7c might be expected. However, since the other mode of oscillation is still possible both on the parallel-line stub (or on the outside of a coaxial stub) so long as the over-all length is such as to make it self-resonant as an antenna both modes will be excited simultaneously on the stub. In the case of the parallel line the equal and opposite currents of the one mode would be superimposed on the unidirectional ones of the other mode; in the coaxial line the current and charge of the one mode is confined to the interior; those of the other mode to the outside. Since the two modes on the parallel-wire stubs exert opposite forces on the outer antennas, the currents in these will be very small.

If a bridge is placed across the center of the $\lambda/2$ stubs as shown in Fig. 7d, the distribution of current may be either like that shown for the left-hand stub or the current may flow through the bridge as shown for the right-hand stub depending upon whether the length of conductor with or without the bridge is more nearly that for resonance. Slight changes in tuning, as for example, an extending or shortening of one side of the stub, will cause the current to shift from the one distribution to the other.

One may conclude from these examples that sections of transmission line are analytically simple and practically effective as coupling or phase-reversing stubs only when they are of such length that they are detuned as antennas. When this is true transmission-line theory alone is adequate. From this one may conclude that a phase-reversing stub should have a high input impedance and that it should be inserted at a point where the antenna without it has a maximum concentration of charge or a minimum current.

If an open stub, which is approximately a quarter wavelength long, is connected at the center of each of the two outer antennas as shown in Fig. 8a a somewhat different problem is presented. In this case the stub is symmetrically placed with respect to the outer antenna so that equal and opposite currents will flow at its terminals. Since in addition the over-all length of the stub is one which completely detunes it from the point of view of an instantaneously unidirectional antenna current on both wires, it may be treated by ordinary line theory. That is, it is entirely equivalent to a very low impedance connected at the center of the antenna. It is easily adjusted to be a pure resistance of the order of magnitude of a few tenths of an ohm[5]. Accordingly the stubs in the outer antennas have practically no effect on the distribution of current in the antenna array at resonance, and this continues to behave much as a single resonant circuit. If each stub is lengthened to a half wavelength as in Fig. 8b it will present an extremely high impedance at the center of each of the outer antennas to a flow of transmission-line currents like those in Fig. 8a. The result is that *this mode of oscillation is completely detuned;* no resonant amplitude

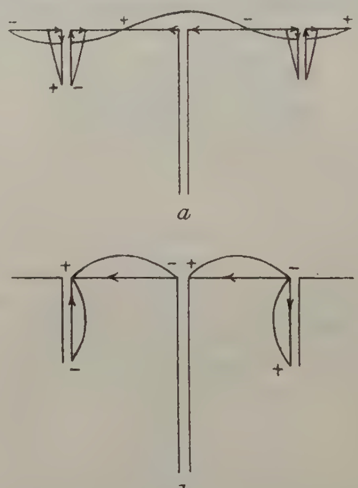


Fig. 8—Antennas and lines coupled in different ways.

of current distributed along the horizontal antennas with a maximum at the input terminals at the center as in Fig. 8a is possible. The input impedance at these terminals is changed from a relatively low to a high value, and if the tuning of the transmission line is readjusted a large resonant amplitude in an entirely different mode as shown in Fig. 8b will be obtained. In effect the "current-fed" or resonant system of Fig. 8a has been transformed into a "voltage-fed" or antiresonant system. A large antenna current flows on each stub as shown instead of a transmission-line current. No resonant currents will flow in the outer parts of the antenna or of the coupling stubs as shown by the absence of current arrows in Fig. 8b. This part of the circuit is actually completely detuned.

If the $\lambda/2$ stubs in Fig. 8b are bridged by conductors at their lower ends, both modes a and b are possible and will be excited simultaneously. The $\lambda/2$ stubs carry an unbalanced current of which a part is a transmission-line-current for mode a with currents in the two wires in opposite directions; a part is an antenna current as in b.

[5] This adjustment as well as the lengths of the several antennas are very critical.

If the $\lambda/4$ stubs in Fig. 8a are bridged by good conductors at their lower ends, a superposition of the modes of Figs. 9a and 9b is to be expected. The parts of the antenna between each stub and the feeding line will swing primarily as in Fig. 9b. Each stub will carry transmission-line currents in directions appropriate for this mode but the outer conductor of each pair will also carry an antenna current so that it will swing with the outermost $\lambda/4$ section of the horizontal antenna just as in Fig. 8a. These outer sections are not detuned in this case as in that of Fig. 8b because they form resonant systems together with the attached parts of the stubs.

The several arrangements discussed above have been described in order to illustrate some of the effects which must be expected in antenna circuits and arrays. An important application which is considered at a later point is the problem of detuning conductors (such as the outside of a coaxial line) which is not intended to be part of an antenna but which may, nevertheless, carry large and undesirable currents. The principle involved is clear from the above illustrations. As in the case of phase-reversing stubs the length must be such that antenna currents are kept negligibly small, so that transmission-line theory is applicable. If this is true it follows directly that a high-impedance stub must be inserted at a point of maximum current (rather than a point of minimum current as for a phase-reversing stub) in order to detune the particular mode. This will reduce the current *due to that particular mode of oscillation* to a minimum. It may, however, make another mode possible.

The Coaxial Colinear Array—Phase-Reversing Sleeves

A very interesting and important alternative arrangement of the colinear array makes use of the inside of a coaxial pipe as feeder, of the inside surfaces of coaxial sleeves (each about a quarter wavelength long and spaced at intervals of about a quarter wavelength) as coupling and phase-reversing stubs and of the outermost surface, consisting partly of the coaxial pipe itself and partly of the coaxial sleeves, as the rather thick units of the antenna. A cross section of this arrangement for two colinear antennas is shown in Fig. 9a with the cross-sectional dimensions enormously enlarged. Actually both the diameter of the pipe and especially the spacing between it and the concentric sleeves are small compared with the wavelength. The complete array consists of the following coupled circuits.

First, there is a coaxial feeder extending from the generator (which is connected to the feeder below J) to K. This line is assumed to be resonant in the figure. (All but a part near the top may be made nonresonant by inserting a suitable matching sleeve.) The currents flowing on the outer surface of the inner conductor and the inner surface of the outer conductor are equal and opposite transmission-line currents. The second coupled circuit is the self-resonant center-driven antenna ABC of which one half is the central conductor AK and the other half the outermost surface BC. (Because the section BC of the antenna is considerably thicker than AK it will have to be shortened considerably more below a quarter wavelength than AK in order to assure self-resonance.) The direction of current and location of maxima at time $t=0$ is indicated by arrows; the location and sign of maximum charge a quarter of a cycle later is also shown. The third coupled circuit is the self-resonant stub of length near $\lambda/4$ with open end at CD and closed end at B. It is formed by the inner surface of the sleeve and the outside of the pipe. The fourth coupled circuit is the antenna consisting of the outer surface DEF. The fifth coupled circuit is the resonant quarter-wavelength sleeve with open end at FG and closed end at E. A sixth circuit consists of the outer surface of the pipe from G to its end below H (including all attached and coupled conductors such as the generator circuit and the earth). This sixth circuit may be resonant or detuned depending upon its length and the arrangement of other attached or coupled circuits. If power is to be radiated primarily from the colinear array made up of the two antennas ABC and DEF, the outside of the feeder line to G must be kept detuned. Methods of achieving this will be considered below. In this case, much as in the case of one of the

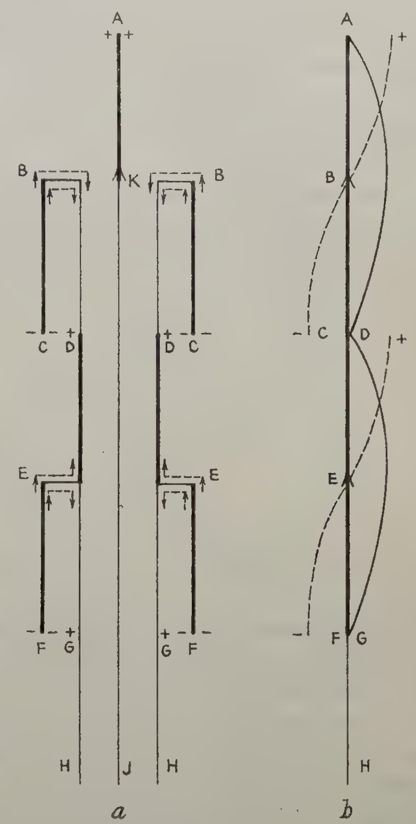


Fig. 9—(a) A coaxial colinear array.
(b) Distribution of maximum current (solid curve) and of charge (dotted curve) a quarter period after the current for an infinitely thin antenna of the same length and construction.

outer antennas of Fig. 4, the antenna DEF is coupled both to antenna ABC and to the section of transmission line with open end at CD. The fields due to these two are opposite in direction, but those due to the sleeve are much greater provided this is adjusted in length to resonance and it is not too small in diameter. This has been assumed in Fig. 9. If this is not true, in particular if the sleeve BC is removed or the open end at CD is closed with a disk, then the retarded forces due to the moving charges in ABC are alone active and the current in DEF is reversed. If the spacing CD is made small by reducing the diameter of the sleeve or if an insulating bead is placed between sleeve and coaxial line at CD, the coupling to the inside of the sleeve may be reduced sufficiently so that the sleeve may also fail to reverse the current.

For optimum performance each of the two antennas and each of the two sleeves must be individually tuned. (Colinear arrays with more than two coupled antennas are readily constructed by attaching additional sleeves below H in Fig. 9.) This cannot be done as accurately in the case of the coaxial colinear array of Fig. 9 as with the structurally less attractive form of Fig. 5a, because the half-length BC (outer surface) of the coaxial antenna cannot be made shorter than the half-length BC (inner surface) of the sleeve. A section of a coaxial line which is open at one end and closed at the other always has a resonant length which differs much less from a quarter wavelength than does the half-length of an antenna of the same or greater outer diameter. Since the antenna and sleeve are closely coupled the adjustment for maximum and as nearly equal currents in the two antennas as possible is not precisely that for self-resonance in each case. It depends therefore on the degree of coupling, and in the absence of an even approximate theoretical treatment it must be determined experimentally.

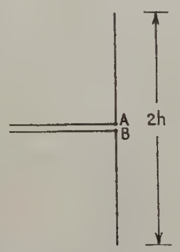


Fig. 10—A symmetrically driven antenna using a parallel-wire line.

The End-Coupled Half-Wave Antenna

A single antenna of length near $\lambda/2$ may be center-driven from a parallel line as in Fig. 10 or from a coaxial line by the arrangement of Fig. 9a for the top antenna ABC. The sleeve EF in this case is not present and the outside of the line below D on the diagram must be detuned by methods to be described later. An antenna of this same length may be coupled to a long, resonant, parallel, or coaxial line, (or a short, resonant, and impedance-transforming section of such a line) using the arrangements of Figs. 11 and 12. Here the antenna is merely the continuation (in the same direction or at right angles) of one of the parallel wires or of the inner conductor of the coaxial line or pipe. Electrically the length AB is a self-resonant antenna which is closely coupled to the resonant two-conductor transmission line and also to the outer surface of the coaxial pipe or to the two wires of the open line acting in parallel as a single conductor. If it is assumed that resonant currents on the outer surface of the coaxial line or on the parallel-wire line acting as a single conductor are minimized by detuning (either by adjustment in over-all length or by other methods to be described below) there remain only two closely coupled circuits. These are the antenna AB and the resonant part of the transmission line. They are coupled by the interaction of forces between charges near the junction point B of the antenna and the line. The degree of coupling depends

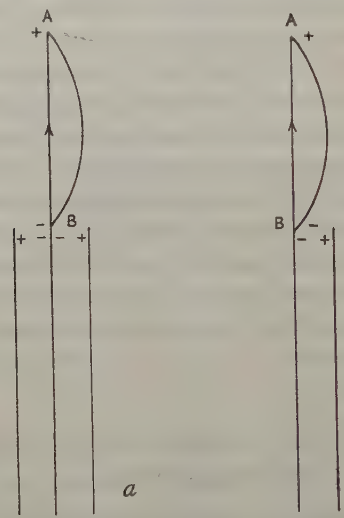


Fig. 11—End-fed antennas.

on the spacing of the two conductors of the line, but unless this is extremely small the coupling will be close, usually greater than critical so that double resonance peaks of current will be observed as the tuning of antenna or line is varied.

If the length of the line is such that the outer surface of the coaxial line or the two wires of the open line treated as a single conductor come into resonance and carry antenna currents, this also becomes a coupled antenna which contributes to the radiation. In the case of the parallel line the resonant antenna currents, which are in the same direction along both conductors, are superimposed upon the equal and opposite transmission-line currents so that the resultant currents are no longer equal and opposite. That is, the parallel line acts simultaneously as both transmission line and antenna as explained above. If the two conductors are enclosed in a metal shield for their entire length the unbalanced or antenna current flows on the outside of the shield just as in the case of the coaxial line.

Unsymmetrical Antennas and Arrays

The symmetry of an antenna or of an array is measured in terms of the geometrical arrangement of the conductors with respect to the two input terminals. In order to define an input impedance for an antenna or array in the conventional low-frequency sense, the two input terminals must be sufficiently close together so that they may be connected as part of a lumped-constant network or, more commonly, to the end of a transmission line the conductors of which are very close together compared with the wavelength. Furthermore, the

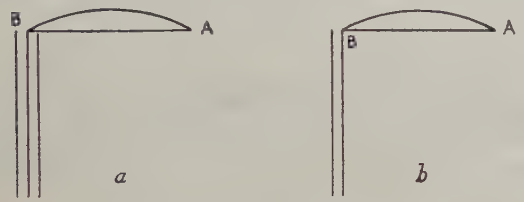


Fig. 12—End-fed antennas.

currents at these two terminals must be equal and opposite if the impedance is to be defined by the relation

$$Z_{AB} = V_{AB}/I_A = V_{AB}/I_B \tag{20}$$

for the two input terminals A and B. In the case of a coaxial line an antenna current flows entirely on the outside of the outer conductor at sufficiently high frequencies so that one can speak of the equal and opposite transmission-line current as a separate current, and it could be measured separately by a meter with its element placed inside the line. On the other hand, with a parallel line an antenna current flows in the same direction on both conductors of the line superimposed on the equal and opposite transmission-line currents. Experimentally there is no way to separate these currents which interact and combine with each other to form a different resultant in each conductor. A long transmission line or a stub section of line can be prevented from carrying a significant antenna current first, if the coefficient of coupling between the line (acting as a single conductor) and the entire array is made vanishingly small, or second, if the coefficient of coupling is not small but the line with all that is attached to it is detuned so that resonant antenna currents are suppressed. In each of Figs. 13 the antenna or array is itself symmetrical and the parallel line is symmetrically placed with respect to its two identical halves. In this case the retarded forces tangent to the line (which are exerted by the moving charges in the two halves of the antenna on the charges in the line) cancel and the mutual impedance between the antenna and the line (treated as a single conductor) vanishes and with it the complex coefficient of coupling. The two wires of the open line thus carry equal and opposite transmission-line currents and no antenna current. In this case the input impedance of the array is easily defined in the conventional way and it could be determined experimentally by substitution or bridge methods or from measurements on the input impedance of the parallel-wire line. The coaxial array of Fig. 9 and the end-fed antennas of Figs. 11 and 12 are unsymmetrical. The charges in the outer surface of the coaxial line (below G in Fig. 9a) experience uncanceled retarded forces from the moving charges in the antennas above, so that the coefficient of coupling does not vanish. In this case the outer surface of the entire coaxial pipe (below G in Fig. 9) must be detuned or it will carry appreciable antenna currents and be a part of the antenna array. The input impedance of the entire antenna array, which is also the terminating impedance of the coaxial line, may be defined in the conventional way because the inner conductors of the line carry equal and opposite currents and charges except very near the open ends. Since a small end correction is in any case always included as part of terminating impedances of more conventional types, the same may be done here. (The equations of the transmission line imply an infinitely long line. If the line is finite, the error, actually made in assuming the constants per unit length of the line to be the same near the ends as far from the ends, is absorbed into the terminating impedance except when the end is open, in which case it leads to a small end correction in length.) In the case of Fig. 11 or Fig. 12 the component of current at the junction point, B, of the line and of the antenna is the component I_z'' in phase with the driving potential difference as previously described[1] for an antenna of nonvanishing radius and of antiresonant length near $\lambda/2$. The approximately equal and opposite current, $-I_z''$, on the inner surface of the outer conductor of the line flows around to the outside of the coaxial line at the upper end, even if this is detuned. For thin antennas it is small. A substitution or bridge method of measuring the impedance *at the terminals* of the antenna would in most cases not be reliable because the outside of the coaxial cable is inevitably a part of the antenna even if detuned, so that it would theoretically have to be detached as a part of the antenna leaving only the inside transmission line. This is, of course, physically impossible. On the other hand, if the constants and the length of the transmission line are accurately known, the input impedance of the line at the generator may be measured by any convenient method and the terminating impedance (which is the input impedance of the antenna) then calculated.

In the unsymmetrical arrangements of Figs. 11b and 12b the two conductors of the parallel line will be completely unbalanced if an appreciable antenna current flows in the same direction in both conductors. It will be somewhat unbalanced even if this is avoided by careful detuning of the line treated as a single antenna because the current at the junction point B of one of the conductors of the line and the antenna does not vanish if the antenna has a physically realizable radius. A component of current,[2] I_z'', inevitably flows into the antenna and this can have no equal and opposite

counterpart on the other parallel wire because this ends and the current must vanish. Accordingly the parallel-wire line in Figs. 11b and 12b is at best slightly unbalanced if the antenna is thin and the line completely detuned as an antenna; at worst, considerably out of balance if the line is not detuned and carries a significant antenna current.

Up to the present, nothing has been said of the possibility of substituting a coaxial line for the parallel line in the completely symmetrical arrangement of Fig. 10. It might perhaps be assumed that the circuit of Fig. 13a using the coaxial line is just as symmetrical as that of Fig. 10 with the parallel-wire line. In so far as the cancellation of *tangential* forces along the outer surface of the coaxial line is concerned, this would certainly be true everywhere except in the proximity of the points A and B (where the antenna is attached) if the two halves of the antenna carried the same current and charges of opposite sign distributed in the same way.

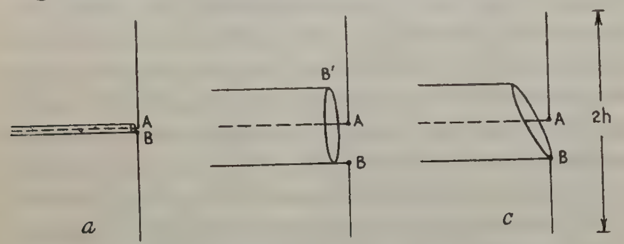


Fig. 13—Center-fed antenna using coaxial line.

The fact is, however, that the distribution of metal and hence the distribution of charge at the end of a coaxial line is always and inevitably unsymmetrical with respect to the two halves of the antenna. Thus, in the enlarged line of Fig. 13b, the periodically charged outer conductor at B' in proximity with the antenna maintains forces on the upper half of the antenna between A and B' which are opposite to the driving forces which are maintained between A and B by the charges on the coaxial line. Similar forces do not act on the lower half of the antenna so that a condition of unbalance obtains. It can be reduced somewhat, but not eliminated, if the end of the coaxial line is cut away near the upper half of the antenna as shown in Fig. 13c. The forces acting on the charges in the two halves of the antenna are still unbalanced. A further condition of unbalance results from the fact that the axis of symmetry for the electromagnetic effects within the coaxial line and at its end is the central conductor, whereas the point of symmetry for the two halves of the antenna is mid-way between the terminals A and B. Another way of expressing this same dissymmetry is in terms of the current which flows out of or into the two halves of the antenna. Thus, that which flows, say, down near A merely continues along the same conductor bent at a right angle, whereas that which flows down must change its rotationally symmetrical distribution along the inner surface of the outer conductor and flow toward a single point B. Accordingly, the adjustment for resonance and the maintenance of similar distributions and amplitudes of current and charge in the two halves of the antenna is somewhat improved if the half of the antenna which is attached to the outer conductor of the coaxial line is made shorter than the other half by about the outer radius of the coaxial line. This also is illustrated in Fig. 13c.

Neither the cutting away of the end of the outer conductor of the coaxial line nor the shortening of the lower half of the antenna is sufficient to assure complete symmetry. In fact the geometrical structure of Fig. 13c is in itself so obviously asymmetrical as to

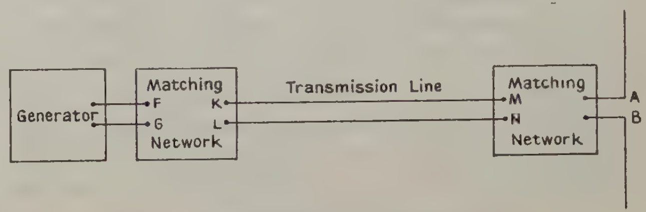


Fig. 14—A typical circuit for an antenna.

make it perfectly clear that the forces acting on charges in the outer surface of the coaxial line in a direction parallel to its axis due to currents and charges in the two halves of the antenna cannot be expected to cancel completely near the end of the line. If they do not cancel, and small tangential forces obtain, current will flow along the highly conducting outer surface of the coaxial line. If the outer surface of the line is not detuned, large resonant amplitudes of current may be built up. In this case the outer surface of the line acts as a coupled antenna in a way similar to that described before.

If the radius of the coaxial line is very small so that the distance AB in Fig. 13a is an extremely small fraction of a wavelength, the dissymmetry is relatively slight. Even so currents due to resonance along the outer surface of the coaxial line may be significant. At very short wavelengths it is not always possible to keep the radius of a coaxial line a negligible fraction of a wavelength because the spacing may then be so small that spark-over can occur. If the distance AB is an appreciable fraction of a wavelength as in Figs. 13b and 13c the dissymmetry is great and relatively large uncanceled forces may be expected to act on the charges in the outer surface of the coaxial line so that this must be kept detuned if resonant amplitudes are to be suppressed.

Transmission-Line Feeders

The connecting circuit between a generator with its associated network and an antenna commonly consists of a transmission line. Such a line is usually of the two-wire, the four-wire, or the coaxial type, and it may be long or short. A typical circuit is shown schematically in Fig. 14. It consists of a generator with output terminals FG, a matching and tuning circuit for the generator with output terminals KL, a transmission line with output terminals MN, and a tuning and matching network for the antenna with output terminals

AB, which are simultaneously the input terminals of the antenna. The most effective over-all transmission of power will be achieved when the impedance looking to the right at FG presents the optimum load for the generator, and when the power losses in the line and in the matching networks are at a minimum. The losses in the matching networks may be kept small by using only circuit elements with low resistance; if sections of line are used these must be kept short. The losses along a transmission line are always least if it carries no antenna current and is terminated in its characteristic impedance Z_c. Methods for satisfying the first condition will be discussed at a later point; the latter condition requires that the impedance looking to the right at MN, viz., Z_{MN}, must be made equal to Z_c. Thus the matching network for the antenna must transform the input impedance Z_{AB} of the antenna into an impedance Z_{MN} looking to the right at MN given by

$$Z_{MN}(\text{right}) = Z_c. \tag{21}$$

If the generator is designed to feed into a load Z_{opt} for optimum performance (such as maximum efficiency or maximum transfer of power), then one must have

$$Z_{FG}(\text{right}) = Z_{\text{opt}}. \tag{22}$$

Accordingly the matching network for the generator must transform the input impedance Z_{KL} (right) of the line at KL into Z_{opt} at FG. If the line is terminated in Z_c as required by (21), then

$$Z_{KL}(\text{right}) = Z_c \tag{23}$$

and the matching network of the generator must transform Z_c looking to the right at KL into Z_{opt} looking to the right at FG. If the generator has been desgined specifically so that

$$Z_{\text{opt}} = Z_c, \tag{24}$$

then the matching network for the generator is unnecessary. At ultra-high frequencies this is usually not possible, and a matching network is required. For most lines $Z_c(=R_c+jX_c)$ is predominantly resistive with X_c very small and negative. Thus for purposes of matching one can write

$$Z_c \doteq R_c; \qquad |X_c| \ll R_c \tag{25}$$

in (21), (23), and (24). Matching networks will not be considered in this paper.

If the conditions (22) and (23) or (24) are fulfilled by a suitably adjusted matching network for the antenna, the transmission line is said to be nonresonant. A nonresonant line is a line that is terminated specifically and only in its characteristic impedance. The primary advantages of the nonresonant line are its relatively low nonresonant potential differences and its low losses. If the line is short, that is, not over a wavelength or two in length, the losses in the line are in any case negligibly small compared with the power transferred to the antenna. Accordingly, it is then relatively unimportant whether the line be nonresonant or not. On the other hand, if the line is long, the losses in the line may become excessive unless the line is made nonresonant.

Even though terminated in its characteristic impedance it may be impossible to make a line nonresonant if the supporting insulators or spacers are incorrectly placed. At long wavelengths supporting insulators are usually separated a negligible fraction of a wavelength both for parallel and coaxial lines. In this case no difficulty is encountered. On the other hand, at extremely high frequencies the spacing of dielectric beads or other supports may be an appreciable fraction of a wavelength. If the beads occur at intervals of slightly less than a half wavelength, successive partial reflections occur at each bead and the effect is cumulative. Accordingly, resonant amplitudes may be built up even though the line is correctly terminated. Exactly the contrary is true if the spacing of the beads is approximately a quarter wavelength. A partial reflection at one bead is then practically canceled by that at the next bead, since the phase difference will be 180 degrees. In this case, therefore, the effect is not cumulative but self-destructive. Transmission lines which are to be made nonresonant at a single, very high frequency should have the insulating supports spaced a quarter wavelength apart if the line is many wavelengths long. If the line is not that long it need not be made nonresonant. An important disadvantage of the nonresonant line is that it does not lend itself readily to multiband operation because it is not possible to provide a single matching network which will terminate the line in its characteristic impedance at more than one or at most two frequencies. For this reason nonresonant lines are seldom used for transmitters operating on a number of frequencies.

For short distances and for multiband operation resonant transmission lines may be used. In this type of operation the antenna-matching network is not required. The impedance Z_{KL} looking to the right at KL is that of the transmission line terminated in the antenna. The condition for optimum performance of the generator then requires the generator matching network to transform Z_{KL} (right) at terminals KL into

$$Z_{FG}(\text{right}) = Z_{\text{opt}} \tag{26a}$$

at terminals FG. The condition

$$X_{FG}(\text{right}) = X_{\text{opt}} \tag{26b}$$

which is contained in (26a) is usually equivalent to tuning the entire circuit to resonance in the matching and tuning network of the generator. In this case, then, the transmission line is also resonant. It is important to bear in mind that a resonant line is not merely any line which is not terminated in its characteristic impedance. It is specifically a line which in conjunction with its termination is tuned exactly to resonance; i.e., its input reactance is the negative of the reactance of the generator. For operation at a single frequency dielectric beads or insulating supports should be placed at current maxima along a resonant line. For operation

at several frequencies they should be so placed that they do not occur near voltage maxima for any of the several frequencies.

Antenna Currents on Lines; Detuning Stubs

Whenever transmission lines are used for purposes which depend on the characteristics of equal and opposite currents as analyzed in conventional line theory it is of primary importance to avoid or at least minimize antenna currents on the line whether this be an open-wire line or a coaxial line. Only if this is accomplished will the line exhibit exclusively those important and extremely useful properties predicted by line theory. These include the transmission of power by a nonresonant or by a resonant line with extremely small

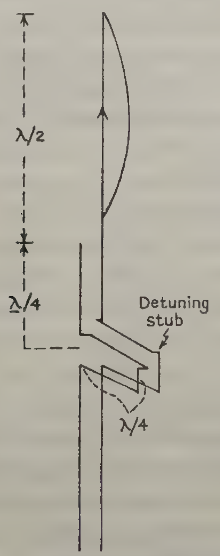


Fig. 15—Detuning stub for a parallel-wire line with unsymmetrical load.

radiation if the spacing of the conductors is a very small fraction of a wavelength; they also include the use of transmission lines for making a large variety of electrical measurements at ultra-high frequencies. In both of these important applications many an engineer and investigator has erroneously come to the conclusion that open-wire lines are unsatisfactory because of large radiation merely because the line was arranged to make an appreciable antenna current possible. Accordingly, it exhibited not only the well-known properties of a transmission line, but superimposed upon these, the equally well-known properties of a linear radiator. In a similar way the large antenna currents which may flow on the outer surface of coaxial lines have caused much trouble when such lines have been used as feeders in many-element arrays.

Antenna currents either in the form of unbalanced currents on parallel-wire lines or currents on the outer surface of coaxial lines are due to two fundamental conditions. First there must be a condition of asymmetry which provides forces acting tangentially along the two conductors which are not quite equal and opposite; second, the line in question must be of such length or so arranged in conjunction with its terminations, that a condition of resonance or at least partial resonance obtains. The complete elimination of antenna currents demands complete symmetry for the terminal impedances and symmetrically applied driving potential differences, as well as over-all detuning of the entire circuit treated as a single conductor. This is not always a simple matter.

If a parallel-wire line is driven by a symmetrically

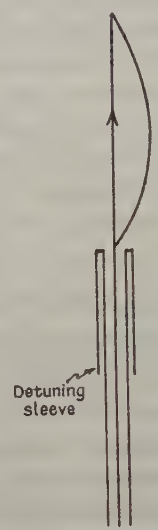


Fig. 16—Detuning sleeve for a coaxial line with an unsymmetrical load consisting of an end-fed antenna.

constructed and symmetrically placed generator[6] either at one end or at any point along the line, and if it is loaded by a symmetrical impedance at the other end, antenna currents can be avoided completely. Since perfect symmetry is difficult to achieve in practice it is always desirable and usually necessary to adjust the over-all length of the line with terminations so that it differs considerably from an integral multiple of a half wavelength. If, on the other hand, a parallel-wire line is driven from a coaxial oscillator of conventional type or from a coaxial line either by direct connection or by means of a coupling unit, complete symmetry is impossible for both the parallel-wire line and the coaxial line since the types of symmetry they require are different. Accordingly appreciable antenna currents are inevitable on either the parallel-wire line, the coaxial line, or both at all frequencies for which the over-all length of either line is resonant in any mode. In such cases the resulting radiation from the antenna current on either or both of the two lines is often incorrectly attributed to radiation from the transmission-line currents of the parallel-wire line. Actually a parallel-wire line the wires of which are separated a small fraction of a wavelength radiates an insignificant amount of power if it does not carry unbalanced currents. And parallel-wire lines can be constructed so that this is the case.

[6] See, for example, Ronold King, "A continuously variable oscillator for parallel line measurements at 100 to 1000 megacycles," *Rev. Sci. Instr.* vol. 11, pp. 270–271; August, 1940.

Unless both the generator and the load of a coaxial line are completely enclosed in metal so that the tangential electromagnetic field at *all* outside points is vanishingly small, the outside surface of the line will always carry at least a small current. This is true in particular if connections are made to the inner and outer conductors and the end is left open. If the over-all length is detuned this current will be so small that radiation due to it will be insignificant just as for the balanced parallel-wire line. If it is not detuned, large antenna currents will flow and radiation will be very appreciable.

Significant radiation from unbalanced currents is automatically avoided in all cases where a transmission line, either open or coaxial, is only a very small fraction of a wavelength from the surface of a highly conducting earth upon which an array of any configuration is erected. The distribution of currents in the highly conducting earth is always such as to lead to a virtual cancellation of the distant field due to the currents on the line.

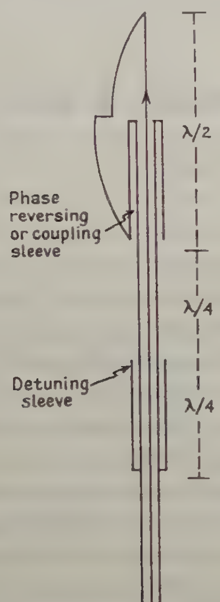


Fig. 17—Detuning sleeve for a coaxial line with an unsymmetrical load consisting of a center-fed coaxial antenna.

If an antenna current along a transmission line cannot be made insignificant by a symmetrical arrangement and by adjusting the over-all length for physical reasons it can always be minimized *for any one mode* by following the method already outlined in connection with Fig. 6b. In this case part of a resonant antenna was detuned by cutting it at a current maximum and connecting the two terminals so obtained to the open end of a high-impedance stub. Any conductor or group of conductors excited in parallel as a single conductor can be detuned in this way with respect to a particular mode by inserting a high-impedance stub at a point where a current maximum would be if the conductor were resonant without the stub. If a line is long and exposed to tangential forces several such stubs may be required. A number of arrangements for detuning parallel-wire lines and coaxial lines which could become resonant as a single conductor are shown in Figs. 15 to 18. In all of these the transverse dimension is very much exaggerated for clarity.

In Fig. 15 a closed-end detuning stub is connected into the parallel line of Fig. 11b. The two sides of the stub are turned to be mutually at right angles so that there is no interaction of the equal and opposite transmission-line currents, whereas to the antenna current, (which flows in the same direction at corresponding points of the two parallel conductors) the two wires of the line are simply in parallel and equivalent to a single one. If the resonant antenna current were to exist on the line, it would have to have a current node at the upper end or junction with the antenna. The first current loop would, therefore, be a quarter wavelength down. If the line is cut at this point and a $\lambda/4$ stub is inserted as shown, the line below the stub is detuned.

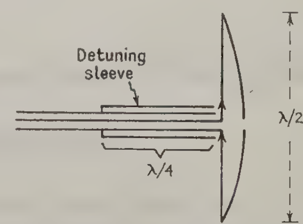


Fig. 18—Detuning sleeve for a coaxial line with an unsymmetrical load consisting of a center-fed antenna at right angles.

In Fig. 16 exactly the same thing is accomplished more simply for a coaxial line. In this case a $\lambda/4$ detuning sleeve is provided with its open end at a quarter wavelength from the top. The sleeve might equally well be moved down a quarter wavelength if the upper end were then left open and the lower end closed.

In Fig. 17 the center-fed antenna uses the outer surface of the phase-reversing or coupling sleeve as its lower half. The inside of this sleeve is a resonant coupled circuit. If the second (lower) sleeve were closed at the top and left open at the bottom it too would be a phase-reversing or coupling sleeve exactly as in the colinear array of Fig. 9a. However, with the open end at the top and the closed end at the bottom as in Fig. 17 it serves to detune the entire circuit consisting of the outside of the coaxial line below the lower end of the upper sleeve. If such a sleeve were placed at a quarter wavelength below the lower coupling sleeve in Fig. 9a the outside of the feeding line of the colinear array would be detuned.

In Fig. 18, the outside of the coaxial line, which is used to center-feed an antenna, is detuned by a suitably placed sleeve. The effects of the partly uncanceled forces acting tangentially along the outside of the line may be summarized roughly by stating that a part of the current from the lower antenna flows along the outside of the coaxial line instead of all along the inside as it would if no dissymmetry existed. Accordingly a resonant current on the outside would have to have its

maximum directly at the end. The open end of a coaxial, $\lambda/4$ detuning sleeve must, consequently, be placed there. If more convenient it could also be placed with its open end a half wavelength from the terminals of the antenna. In all cases where an antenna is fed from a flexible coaxial line (as in Fig. 18) a detuning stub is essential. For if the cable is moved or coiled, its electrical length is altered and resonance may be established in one position and not in another. This is a most undesirable condition since the impedance at the terminals of the antenna as well as the power radiated may become a function of the manner in which the feeding cable is coiled. Detuning sleeves should be considerably larger in diameter than the coaxial line to provide adequate coupling.

Experimental Verification

The discussion of coupled transmission lines and antennas has been nonmathematical because a complete analytical treatment of the difficult coupling problems is unavailable. It has, nevertheless, been based on general electromagnetic principles in so far as a qualitative application of these was possible. In addition, practically all phenomena discussed above were verified in detail experimentally both in the laboratory and on the lecture table. Special equipment was constructed for this purpose using a wavelength of 1 meter. Space does not permit its description at this point.[7]

[7] A detailed description of this and other demonstration and laboratory equipment for ultra-high-frequency and microwave work is to be published in *Electronics*.

Radio Production for the Armed Forces*

STANFORD C. HOOPER†, ASSOCIATE, I.R.E.

After one year in World War II we find that almost the entire radio facilities of the United States are now geared to production of radio and underwater sound apparatus for the Armed Forces and Merchant Marine. In addition, many plants and laboratories not previously in this branch are producing this equipment. There are over 500 plants engaged, with approximately 200,000 employees. This includes the subcontractors. It is one of the largest and most important of war industries.

Communications within the Armed Forces, detection and location of the enemy, identification, and radio control are essential to nearly every mobile war unit, in some form. Rapidity of movement of forces in the air, on the ground, on and under the sea, makes success impossible without instantaneous, secure, and positive communication.

A dramatic illustration of this was provided during our invasion of North Africa. The deadly accuracy of firing by one of our battleships which destroyed the *Jean Bart*, at Casablanca, was made possible by radio communication. As the first blast from our big guns, twenty-six miles away, struck the resisting French battleship, an observation plane flashed back the word of a direct hit on the deck, a damaging but not a fatal blow. A slight change in elevation was signaled for. The next salvo struck the side of the ship at the water line, smashing her hull beyond repair. Radio directed and reported the destruction.

Production, enrollment of personnel, and training in the radio industry are proceeding according to well-formulated plans and are steadily but surely reaching the point where requirements will be satisfied, at least until the demands for increased production are again raised, which no doubt will be very soon. Every time a ship goes down, demands for production are increased. We must not only replace that lost equipment, we must strive to duplicate it in order to outstrip the Axis.

In Berlin today, in Rome, and in Tokio, there are groups just like yours, which are working to outproduce us. Just as the Navy must outfight the enemy, ship for ship, and plane for plane, so must you in this industry, outthink, outplan, and outproduce the Axis competition in the production of radio facilities.

Production is not a simple thing. It includes design; and many designs for each order of apparatus are made and discarded before the final one is decided upon. It includes manufacturing drawings which, in the case of a complete ship, may mean one hundred and fifty tons of blueprints, alone. It includes the procurement of raw materials, materials which must be transported from far-distant sources of supply, through dangerous waters and at great hazard, and new materials to replace those of which the enemy has deprived us by seizure or occupation. Production includes manufacture of parts, their assembly and test, thorough inspection during and after the process of manufacturing and assembly, packing, shipment, and delivery to destination. All along the line expediters are engaged night and day in the work of hastening the delivery of materials and parts to keep production lines moving. Then the finished product goes to war, taking its place with the Armed Services for the final test.

It is amazing to see the complete change from peacetime to wartime activity in the radio industry in the United States. A year ago these same plants were producing broadcast receivers at the rate of 4,000,000 a

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† Rear Admiral, United States Navy (Retired), 72 Spier Drive, South Orange, New Jersey.

year, and a thousand other products for civilian use. Now they comprise one of the largest co-ordinated production lines for the war effort in the world, under the co-ordinated direction of the Army, Navy, and War Production Board.

It is hard for the majority of radio engineers to appreciate the need for the complex specifications governing design of service apparatus. For thirty years we have needed improvements which we are now getting. The experience of the Armed Forces over thirty years has demonstrated the need for them. Now, under the compulsion of war, we are getting them and getting performance which has never before been attained. These new specifications reflect the demand for perfect performance; perfect reception by planes flying at twenty thousand feet, battling ice and sleet, as well as the enemy; perfect reception by pitching tanks, hurdling debris, and jolting through shell holes in the heat of the African deserts; perfect reception for all our mobile equipment, whether it be in the Battle of Midway, the Aleutians, or the green hell of steaming jungles in the Solomons.

These specifications call for equipment that must stand up with full efficiency under all conditions—tropical and Arctic temperatures, rapid changes in altitude, varying humidities, salt spray, hot sun, and desert sands. It must be unaffected by the motion of motorized units, ships and aircraft, and the jar and vibration due to gunfire and shell impact. It must be fireproof, especially from the instantaneous hot flame which follows a bomb explosion or proximity to hot metal surfaces. It must carry on during severe icing and snow conditions. It must be rugged to withstand mishandling and operation by inexperienced personnel, and jars due to handling in transit. It must be designed to compromise ruggedness and extreme sensitivity. It must be capable of being operated adjacent to various other transmitters and receivers through the roar of battle, through electrical and other noises of ships and planes, and radio jamming. The radiation from tubes must not divulge presence to an enemy. It must be flexible in frequency shifting and power variation in order that shifts from one command or information channel to any other may be accomplished as required, and instantaneously. It must be constructed for installation in most limited spaces, with minimum weights, and convenient for operation. It must be instantaneously ready for operation at all times, exactly on the prescribed frequency, and accessible for adjustment and quick repair. Danger of accident due to electric shock to personnel must be prevented. These are but a few examples to show the need of specifications more elaborate than those governing design of commercial equipment.

Those who have not served in the Army or Navy may find it difficult to appreciate why such strict standardization is required, and why so many spare parts. Standardization is necessary in order that radio operators anywhere may be able to communicate with one another on the prescribed channels and within prescribed methods of operation, under constantly shifting conditions of location, changes in combination of units within the various forces, change in commanders, and many military factors involved.

Also, standardization is vitally necessary because of the constant shifting of units from one location to another. Ships, planes, and tanks must be able to obtain parts from the various widely separated shore bases and constantly changing supply vessels afloat, and unless there are plenty of spares, and all standard, there would be no guarantee that the stocks would contain the exact spares needed by the particular unit. Therefore, it is necessary to keep stocks of all spare parts in practically every base afloat and on land, all over the world. This should be borne in mind by the research and design engineers so that in their designs they will adhere to a minimum of standard parts such as condensers, resistors, meters, tubes, nuts, and bolts.

The public hears much about production activities, but little about what the research engineers are doing. It might be of assistance to the enemy if the magnitude of research work were discussed, but I can say that great strides are being made in research, and co-ordination exists between the activities. This story will be of tremendous interest after the war and the results will have a far-reaching effect on the improvements which will be adaptable to public uses. This is always the case following a great war.

If the war lasts many years, which we pray it will not, the research work now being carried on may prove to be the margin for victory. Much improvement in the equipment is already apparent due to the research engineers. If we were able to predict the length of the war, then it would be possible to say exactly the date research should cease, a date chosen when all equipment design should be frozen for the remainder of the war and a portion of those engaged in this field could be made available to assist in production.

In the first World War our radio and sound production just about equalled one week's production in World War II. This shows the great advance in appreciation and use of radio, with increased speed due to airplanes and fast ships, which has now changed the movements of warfare from hours to seconds. Battles are won and lost on the strength of communications. Some of Rommel's earlier successes in Africa were due not so much to the numbers of his tanks as to the superiority of his communications. Strength of forces, their placement, and the "edge" gives victory. The "edge" is given to the commands which keep the initiative and can first detect changes in the strategical situation. Radio and sound are the instruments used for this quick service.

I want to say, for the Armed Forces, that we are thoroughly cognizant of the production situation as it exists in radio and underwater sound. All of us, from the Commanders-in-Chief down, know the apparatus you make because we use it constantly whether in the task

forces or engrossed in the business or inspection end relating to its procurement. Discussion concerning the comparative merits of the various types and makes of apparatus is a regular "wardroom-mess" discussion these days.

I take this opportunity to express our sincere appreciation of the work of the engineers who have been brilliant, farsighted, and courageous. The production personnel deserves equal praise for their devotion to duty, and long hours in planning and putting into operation the radio industry's part in the united mass production effort of this war. Together, let us all march forward in the electronics victory production parade.

Standard-Frequency Broadcast Service of National Bureau of Standards, United States of America*

THE National Bureau of Standards, as stated in an announcement of August 1, 1943, broadcasts standard frequencies and related services from its radio station WWV, at Beltsville, Md., near Washington, D. C. The service has been improved and extended, a new transmitting station has been built, 10-kilowatt radio transmitters installed, and additional frequencies and voice announcements added. The services include: (1) standard radio frequencies, (2) standard time intervals accurately synchronized with basic time signals, (3) standard audio frequencies, (4) standard musical pitch, 440 cycles per second, corresponding to A above middle C.

The standard-frequency broadcast service makes widely available the national standard of frequency, which is of value in scientific and other measurements requiring an accurate frequency. Any desired frequency may be measured in terms of any one of the standard frequencies, either audio or radio. This may be done by the aid of harmonics and beats, with one or more auxiliary oscillators.

The service is continuous at all times day and night. The standard radio frequencies are:

5 megacycles (=5000 kilocycles=5,000,000 cycles) per second, broadcast continuously.

10 megacycles (=10,000 kilocycles=10,000,000 cycles) per second, broadcast continuously.

15 megacycles (=15,000 kilocycles=15,000,000 cycles) per second, broadcast continuously in the daytime only (i.e., day at Washington, D. C.).

All the radio frequencies carry two audio frequencies at the same time, 440 cycles per second and 4000 cycles per second; the former is the standard musical pitch and the latter is a useful standard audio frequency. In addition there is a pulse every second, heard as a faint tick each second when listening to the broadcast. The pulses last 0.005 second; they may be used for accurate time signals, and their 1-second spacing provides an accurate time interval for purposes of physical measurements.

The audio frequencies are interrupted precisely on the hour and each five minutes thereafter; after an interval of precisely one minute they are resumed. This one-minute interval is provided in order to give the station announcement and to afford an interval for the checking of radio-frequency measurements free from the presence of the audio frequencies. The announcement is the station call letters (WWV) in telegraphic code (dots and dashes) except at the hour and half hour when the announcement is given by voice.

The accuracy of all the frequencies, radio and audio, as transmitted, is better than a part in 10,000,000. Transmission effects in the medium (Doppler effect, etc.) may result in slight fluctuations in the audio frequencies as received at a particular place; the average frequency received is however as accurate as that transmitted. The time interval marked by the pulse every second is accurate to 0.00001 second. The 1-minute, 4-minute, and 5-minute intervals, synchronized with the seconds pulses and marked by the beginning and ending of the periods when the audio frequencies are off, are accurate to a part in 10,000,000. The beginnings of the periods when the audio frequencies are off are so synchronized with the basic time service of the United States Naval Observatory that they mark accurately the hour and the successive 5-minute periods.

Of the radio frequencies on the air at a given time, the lowest provides service to short distances, and the highest to great distances. For example, during a winter day good service is given on 5 megacycles at distances from 0 to about 1000 miles, 10 megacycles from about 600 to 3000 miles, and 15 megacycles from about 1000 to 6000 miles. Except for a certain period at night, within a few hundred miles of the station, reliable reception is in general possible at all times throughout the United States and the North Atlantic Ocean, and fair reception over most of the world.

Information on how to receive and utilize the service is given in the Bureau's Letter Circular, "Methods of

* Decimal classification: R555. Original manuscript received by the Institute, August 5, 1943.

Using Standard Frequencies Broadcast by Radio," obtainable on request. The Bureau welcomes reports of difficulties, methods of use, or special applications of the service. Correspondence should be addressed National Bureau of Standards, Washington, D. C.

Correction

S. P. Chakravarti, whose paper, "A Note on Field Strength of Delhi 3 and Delhi 4 at Calcutta During the Solar Eclipse of September 21, 1941," appeared on pages 269 to 271 in the June, 1943, issue of the PROCEEDINGS, has brought to the attention of the editors the following correction.

The first sentence of the paper reads as follows: "Field strengths of Delhi 3 (5 kilowatts aerial power and 19.62 meters day wavelength) and Delhi 4 (5 kilowatts aerial power and 25.26 meters wavelength) were simultaneously measured at Calcutta," etc. This should read: "Field strengths of Delhi 3 (5 kilowatts aerial power and 19.62 meters day wavelength) and Delhi 4 (10 kilowatts aerial power and 25.36 meters wavelength) were measured at Calcutta," etc.

Institute News and Radio Notes

Electronics

The radio-and-electronic field has been analyzed in a discussion by Carl J. Madsen, electronics engineer in the industry engineering department of the Westinghouse Electric and Manufacturing Company. Mr. Madsen has been an Associate of the Institute since 1928. Extracts from his analysis of the field follow:

There are probably as many definitions of the word electronics as there are individuals associated with its application. One definition is: electronics is the application of devices in which the flow of free electrons are made to perform such numerous functions or duties, as to rectify, amplify, generate, control, convert light into current and current into light. The necessary tubes may take the form of diodes, triodes, ignitrons, pentodes, beam power tubes, thyratrons, phanatrons, kenatrons, phototubes, cathode-ray tubes and so on. A brief study of the application of these hundreds of types of tubes shows that the science of electronics is not new. It also shows that the future possibilities are extremely important in that hardly a week passes without the development of new tube types or new combinations of tubes and circuits to perform new functions.

The developments of electronics can best be explained by breaking down the applications into these nine classifications: rectification, inversion, high-frequency heating, communications, measurements, control, inspection and sorting, precipitation, and radiation.

There are two fields which have been important in past years in rectification. The first is power rectification. The second, high-voltage rectification, is important in that it is frequently employed as a means to an end in many other types of electronic equipment, such as, power supply for high-frequency oscillators, communication equipment, measuring equipment, and broadcast receivers and transmitters. Rectification will have an important position in most of the electronic devices which may be developed in the future because of the nature of electronic tubes.

High-frequency heating was the subject of much experimental work more than ten years ago. Experiments were conducted at that time in the heating of various materials such as plastics, bonds, food, metals, cements, and for the extermination of bugs and larvae. Excessive cost caused by misapplication or misunderstanding of some of the limiting factors proved some of these applications unpractical. However, a number of these applications have been brought to the forefront recently and will undoubtedly become increasingly widespread in their application in the future.

Some of the present applications employing dielectric heating are the bonding of plywood and the heating and curing of plastic materials. Dielectric heating has its important application where thick sections of plywood, thermoplastic, or thermosetting materials are involved. The development of thermosetting bonding materials now permit fabrication of thick sections of plywood in the matter of three to five minutes. Hours were required with steam or other form of heat. The same is true in the heating and curing of plastics. In addition to speed, the electronic method does a more thorough and uniform job. High-frequency heating is bound to have an important position in the future in both the plastic and plywood industries.

Another important phase of high-frequency induction heating is used in the heating of metals. With this method, faster and more uniform heat treating, annealing, brazing, welding, soldering, and tempering have been made possible. In some processes, time has been reduced from 2 minutes to 5 seconds. It is possible, by proper choice of frequency and equipment, to case-harden desired surface of mechanical parts, such as gears and shafts, and leave the base metal tough and malleable.

The conservation of tin in making tin plate was made possible by the application of induction heating. Tin plate is produced at speeds approaching 1000 feet per minute. Here high-frequency power in a single plant equals the total power of all our conventional broadcast stations, and the total installations will soon be over 2½ times this figure. It is only a year ago that the early experiments were made on tin plate.

To date, induction heating has been limited primarily to those particular applications important to our war program. After the war is over and the limiting restrictions of supply have been removed, hundreds of new high-frequency heating applications important to the steel, aluminum, tool, and general manufacturing industries will be found for electronics.

In the field of communication considerable prophecy has already been made by many leading authorities on frequency modulation, television, and broadcasting. Important developments in the past year or so will lead to vast expansion in the number of frequency-modulation and television sets. Even in our present broadcast field the trend of ever-increasing power leads us to predict the building of "superpower" broadcast stations of 750,000 to 1,000,000 watts output.

Carrier-current transmission, a less known phase of communication, has found increasingly wide application in the past few years. Its use in connection with protective relaying has permitted the capacity of our present power lines to be increased over 50 per cent, with a comparable saving in vital copper and other critical materials. Telemetering, or the remote indication of circuit and plant loading to a central dispatcher in the power-distribution system, is another application of carrier-current developments. This application facilitates the economic operation of power-generating systems. As the power needs of the country grow, this industry will find some of the wartime developments in electronics ready to assist in their problems.

In the field of electronic devices for making measurements, a number of developments have been made in the past few years. These developments include dynetric balancing, the electron-mass spectrometer, cathode ray, stroboglow, micrometers and many others. Dynetric balancing is highly important today as it measures and locates the position of off-balance components of all types of rotating parts from the tiny aircraft instrument gyros weighing less than a quarter of a pound, to the massive marine gears weighing over 80 tons. In the conventional sizes, rotors weighing from ½ pound to 100 pounds are often balanced in less than 15 seconds. Former methods required more than one hour. Off-balance components producing vibrations of as low as three thousandths of an inch can be accurately measured electronically and the position for a balancing weight located within two degrees. This development will be of increasing importance in years to come in helping to build longer-life machines with quieter and more dependable operation.

Many types of electronic equipment have been developed for use in the inspection and sorting of products in a diversified group of industries. Industrial X ray, for instance, is at present applied for the inspection of armor plate, welds, airplane parts, and other equipment, spotting certain defects which might otherwise escape notice. In peacetime the advantages of this type of inspection may result in safer, lighter automobiles, trucks, and planes as unnecessary safety factors for unseen defects can be avoided.

Photoelectric devices in applications such as pinhole detectors permit the rapid inspection and automatic sorting of prime and defective sheets. Defects which often escape visual inspection are spotted faster than the combined work of a dozen employes. Simple operations such as high-speed counting and the control of conveyor lines are applications which are in operation today, but which undoubtedly will be extended to many other industries in the near future.

It is not possible to anticipate all the future developments made possible by electronics. The possibilities are beyond imagination. However, many applications exist where electronics can do the work better than other types of equipment, do some things not possible in any other way. But electronics cannot do everything. Many possible applications are not economical or practicable. With electronics now on the tip of every tongue it is essential that every experienced electronics engineer weigh carefully every request, every possible application to avoid misapplications which may lead to delay in acceptance on jobs it can do well.

The vast production capacity developed to meet present wartime requirements will place electronics in the position to serve the needs of the postwar world, but our development effort and production facilities must be used wisely. It is important that we "keep our feet on the ground" lest the electronic field again be smothered by adverse publicity caused by misapplication.

Board of Directors

The regular meeting of the Board of Directors took place on September 8, 1943, and was attended by L. P. Wheeler, president; F. S. Barton, vice president; S. L. Bailey, E. F. Carter, I. S. Coggeshall, H. T. Friis, Alfred N. Goldsmith, editor; O. B. Hanson, R. A. Heising, treasurer; F. B. Llewellyn, Haraden Pratt, secretary; G. T. Royden (guest), B. J. Thompson, H. M. Turner, H. A. Wheeler, W. C. White, and W. B. Cowilich, assistant secretary.

These applications for membership were approved: for transfer to Member grade, Stanford Goldman, S. L. Seaton, and G. C. Sziklai; for admission to Member grade, D. D. Cole, P. G. Forsyth, P. M. Gunzbourg, G. N. Hancock, M. K. Toeppen, and Marc Ziegler; Associate grade, 238; Student grade, 156; and, Junior grade, 6.

The report of the Tellers Committee relative to counting and checking of the Constitutional-amendment ballots, dated June 30, 1943, was accepted and the Constitutional amendments declared adopted.

President Wheeler called attention to a letter concerning a proposed change in name of the Institute, and to the recent decision of the War Production Board on the Institute's appeal for relief from that agency's Paper Limitation Order L-244.

Secretary Pratt, as chairman of the Committee on the Radio Technical Planning Association, reviewed the developments and announced that a meeting of invited sponsors would take place on September 15, 1943.

Unanimous approval was granted to holding the Winter Conference and Annual Meeting during January, 1944, in New York City, and to a program of utmost brevity consistent with the presentation of papers of importance to the present emergency. Dr. B. E. Shackelford was appointed chairman of the committee on arrangements.

It was decided to present the Institute's Medal of Honor, Morris Liebmann Memorial Prize, and Fellowship Awards for 1943 at an evening session of the forthcoming Winter Conference.

A committee consisting of Treasurer Heising, chairman; President Wheeler, Secretary Pratt, Editor Goldsmith, and Assistant Secretary Cowilich was appointed to investigate other office quarters which would allow for the expansion of Institute activities.

The report on the Institute investments presented by Treasurer Heising, as chairman of the Investment Committee, was accepted and the recommendations approved.

Proposed amendments to the Bylaws, submitted by Treasurer Heising, were received and tabled for further consideration.

On recommendation of the Executive Committee, approval was given to continuing Institute Representatives on Other Bodies, indicated below:

- American Documentation Institute: J. H. Dellinger
- Council of the American Association for the Advancement of Science: J. C. Jensen
- Joint Co-ordination Committee on Radio Reception of the E.E.I., N.E.M.A., and R.M.A.: C. E. Brigham
- National Research Council, Division of Engineering and Research: F. E. Terman
- U.R.S.I. (International Scientific Radio Union) Executive Committee: C. M. Jansky, Jr.
- U. S. National Committee, Advisers on Electrical Measuring Instruments: Melville Eastham and Harold Olesen
- U. S. National Committee, Advisers on Symbols: L. E. Whittemore and J. W. Horton
- ASA Standards Council: Alfred N. Goldsmith (H. P. Westman, alternate)
- ASA Electrical Standards Committee: H. M. Turner (H. P. Westman, alternate)
- ASA Sectional Committee on Acoustical Measurements and Terminology: E. D. Cook and H. F. Olson
- ASA Sectional Committee on Definitions of Electrical Terms: Haraden Pratt
- ASA Subcommittee on Vacuum Tubes: B. E. Shackelford
- ASA Sectional Committee on Electric and Magnetic Magnitudes and Units: J. H. Dellinger
- ASA Sectional Committee on Electrical Installations on Shipboard: I. F. Byrnes
- ASA Sectional Committee on Electrical Measuring Instruments: Wilson Aull
- ASA Sectional Committee on Graphical Symbols and Abbreviations for Use on Drawings: Austin Bailey (H. P. Westman, alternate)
- ASA Subcommittee on Communication Symbols: H. M. Turner
- ASA Sectional Committee on Letter Symbols and Abbreviations for Science and Engineering: H. M. Turner
- ASA Subcommittee on Letter Symbols for Radio Use: H. M. Turner
- ASA Sectional Committee on National Electrical Safety Code, Subcommittee on Article 810, Radio Broadcast Reception Equipment: E. T. Dickey (Virgil M. Graham, alternate)
- ASA Sectional Committee on Preferred Numbers: A. F. Van Dyck
- ASA Sectional Committee on Radio: Alfred N. Goldsmith, chairman; Haraden Pratt, and L. E. Whittemore
- ASA Sectional Committee on Radio-Electrical Co-ordination: J. V. L. Hogan, C. M. Jansky, Jr., and L. E. Whittemore
- ASA Sectional Committee on Specifications for Dry Cells and Batteries: H. M. Turner
- ASA Sectional Committee on Standards for Drawings and Drafting Room Practices: Austin Bailey
- ASA Committee on Vacuum Tubes for Industrial Purposes: B. E. Shackelford
- ASA War Committee on Radio: Alfred N. Goldsmith*

* Also Chairman of Its Subcommittee on Insulating Material Specifications for the Military Services.

The appointment to the Electronics Committee of these three subcommittee chairmen took place:

- A. M. Glover, chairman, Subcommittee on Photoelectric Devices
- D. E. Marshall, chairman, Subcommittee on Gas-Filled Tubes
- A. L. Samuel, chairman, Subcommittee on Advanced Developments

President Wheeler reviewed the need for expanding the scope of Institute activities and Dr. Llewellyn was requested to prepare a report on the subject for the next meeting.

The distribution of the Temporary Facsimile Test Standards to the entire membership was approved.

Executive Committee

The Executive Committee met on September 7, 1943, and those present were L. P. Wheeler, president; Alfred N. Goldsmith, editor; R. A. Heising, treasurer; F. B. Llewellyn, Haraden Pratt, secretary; H. A. Wheeler, and W. B. Cowilich, assistant secretary.

The following applications for membership were approved for confirming action by the Board of Directors: transfer to Member grade, Stanford Goldman, S. L. Seaton, and G. C. Sziklai; admission to Member grade, D. D. Cole, P. G. Forsyth, P. M. Gunzbourg, G. N. Hancock, M. K. Toeppen, and Marc Ziegler; Associate grade, 238; Student grade, 156; and, Junior grade, 6.

Action was taken on certain office matters, including overtime work, which were presented by Assistant Secretary Cowilich.

The subject of office quarters was discussed and followed by a recommendation to the Board of Directors.

Matters pertaining to the Buenos Aires Section were given further consideration.

Editor Goldsmith reported on the decision of the War Production Board relative to the Institute's supplementary appeal for relief in behalf of the PROCEEDINGS, from that agency's Paper Limitation Order L-244.

It was indicated by Editor Goldsmith that the number of papers on hand are sufficient for the next few issues of the PROCEEDINGS and that the situation can be considered good.

Attention was called by Editor Goldsmith to the Audit Bureau of Circulations statement covering the first six-month period of 1943, which had been completed by the staff and which had recently been released by that Bureau.

Mr. H. A. Wheeler, as chairman of the Standards Committee, gave a progress report on the Standards on Symbols and the Temporary Facsimile Standards.

Following a review by Dr. Llewellyn, the list of Institute Representatives on Other Bodies was revised and in that form recommended to the Board of Directors.

Several matters pertaining to the holding of a Conference and Annual Meeting were considered and followed by recommendations to the Board of Directors.

The holding of a meeting of the Sections Committee meeting was given consideration.

Recommendations were made to the Board of Directors on additional personnel for the Electronics Committee.

At the suggestion of President Wheeler, the desirability of expanding the scope of Institute activities was discussed.

Treasurer Heising described the report on Institute investments and read a proposed Bylaws amendment relative to the establishment of a permanent Investment Committee, in advance of the submission of these matters to the Board of Directors.

Government Radio Official Commended for Long Service

The Federal Communications Commission made public on August 24, 1943, a letter to Mr. William D. Terrell, Chief of the FCC's Field Division, Engineering Department, who terminated 40 years of government service when he retired August 31, 1943.

Born August 10, 1871, in Golansville, Virginia, Mr. Terrell is one of the outstanding pioneers in communications science. As Chief of the Radio Division of the Department of Commerce from 1915 to 1932 he contributed perhaps more than any other government official to the growth of broadcasting and high-frequency communications. In 1932 Mr. Terrell became Chief of the Division of Field Operations of the Federal Radio Commission and in 1934 took over the direction of the Field Division of the newly-created FCC. The letter follows:

Mr. William D. Terrell
Chief, Field Division
Engineering Department
Federal Communications Commission
Washington, D. C.

Dear Mr. Terrell:

On the occasion of your voluntary retirement from government service August 31, 1943, may I convey to you on behalf of the Commission and its staff, as well as personally, our sincere best wishes and our hope that you will continue to enjoy for many years to come health, happiness, and the satisfaction of important work well done. We know that the friendships cemented during our association with you will endure, and that you will continue to hold the respect of all concerned with radio which you have earned during your forty years of meritorious service to your government.

In 1911, when you became the first United States Radio Inspector, you had already had twenty-two years of pioneer communications experience including eight years of government service. Thereafter, as Chief of the Radio Division of the Department of Commerce, you contributed more than any other government official toward the early growth of broadcasting and of high-frequency communication. Since 1932, as Chief of the Division of Field Operations of the Federal Radio Commission, and as Chief of the Field Division of the Federal Communications Commission, you have devoted yourself unremittingly and unsparingly to the duties of your office.

We especially wish to thank you for your last two years on active duty, undertaken at our request and with the approval of the President after you had passed seventy, the statutory age of retirement for Federal employees, thus giving us the benefit of your expert advice and assistance during the most difficult period of adjustment to war conditions when your help was urgently needed.

As tokens of your accomplishment and of the esteem in which you are held in your profession, you were elected a Fellow of the Institute of Radio Engineers in 1929 and made an honorary member of the Veteran Wireless Operators Association. You have represented this Government with distinction at many national and international meetings, including the International Radiotelegraph Conference, London, 1912; National Broadcast Conferences called by the Secretary of Commerce, 1922, 1923, 1924, and 1925; International Telegraph Conference, Paris, 1925; International Radio Conference, Washington, 1927; Safety of Life at Sea Conference, London, and European Broadcasting Conference, Prague, 1929. In all these lines of duty, you have brought credit to yourself and the government.

Not the least of your services has been the selection and training of younger men who will now carry on the tradition of competence and integrity which you have established, and who will seek to maintain the high standards you have set. I know these men join with the Commissioners in appreciation and cordial best wishes.

BY ORDER OF THE COMMISSION
James Lawrence Fly
Chairman

THE WHITE HOUSE

Washington, D. C.
August 31, 1943

Dear Mr. Terrell:

I take the occasion of your retirement from Federal service to convey to you my thanks and gratitude for the forty years service in the field of governmental radio services.

You can well be proud of the record you have made.

Very sincerely yours,
Franklin D. Roosevelt

Mr. W. D. Terrell
4764 24th Read North
Arlington, Virginia

David Grimes

David Grimes, vice president in charge of engineering for Philco Corporation, who was abroad on a special war mission, was killed on September 4, 1943, when the transport plane in which he was traveling, crashed into a mountain in Northern Ireland. He was 47 years of age.

Mr. Grimes was born on May 28, 1896, at Minneapolis, Minnesota. Following his graduation from the University of Minnesota, he served in the last war as chief radio officer at Kelly Field, Texas, when the use of radio communications in warfare was just beginning to assume importance. From June to December, 1918, he was Signal Officer attached to the British Air Forces at Aldershot and Littlehampton, England.

After the war, Mr. Grimes joined the American Telephone and Telegraph Company as a research engineer in telephony. In 1922 he established his own engineering organization to do research work on a consulting basis for a number of different companies. It was during this period that he invented the famous "Grimes Inverse Duplex circuit" that was used by many early radio amateurs in home-made receivers. From 1930 until 1934, he was license engineer with the Radio Corporation of America.

Mr. Grimes joined Philco in 1934 as engineer in charge of home-radio-set research and engineering, and continued in that capacity until 1939, when he was named chief engineer. In 1942 he was elected vice president in charge of engineering.

Under Mr. Grimes's direction, the Philco research and engineering department was greatly expanded and strengthened. His fellow officials regret that Mr. Grimes will not see the fruition of many of his advanced ideas.

Mr. Grimes contributed in substantial measure to the development of the radio industry and did much to prepare the way for the general introduction of television after the war. Under his direction, his organization established one of the first successful television relay systems, which was put into operation between New York and Philadelphia in October, 1941 It was his belief that a network of similar relay links which beamed television programs through the air from one station to another 25 to 40 miles apart would make it possible to develop a nation-wide television service in a relatively short time. Mr. Grimes joined the Institute of Radio Engineers as an Associate in 1920 and transferred to Member Grade in 1928.

I.R.E. People

Nelson P. Case

Nelson P. Case, Member of the I.R.E., has been appointed director of the newly created engineering, design, and development division of Hamilton Radio Corporation of New York. The program of his work will include the design and development of household radio models for postwar use.

Mr. Case was associated for approximately thirteen years with the Hazeltine Electronics Corporation. He has served on several I.R.E. committees and, until the war started, was chairman of the receiver section of the R.M.A. Engineering Department.

Palmer M. Craig

Palmer M. Craig, for the past two years chief engineer in charge of radio communications equipment development, has been named chief engineer of the radio division

of Philco Corporation. Mr. Craig joined the Philco Research Laboratories in 1933 as a radio engineer and assisted in the development of high-fidelity reception, automobile radios, and the remote-control radio receiving sets. He was appointed engineer in charge of console radios in 1938 and, even prior to the attack on Pearl Harbor, was active in the development of military radio equipment.

He was graduated from the University of Delaware in the class of 1927 with the degree of B.S. in electrical engineering and was formerly associated with the Westinghouse Electric and Manufacturing Company. He has been an Associate Member of the Institute since 1925.

H. H. Friend

H. H. Friend, who until recently was associated with Scintilla Magneto Division of the Bendix Aviation Corporation, is now development engineer of electronics, airplane division department, of the newly formed development division, Curtiss-Wright Corporation at Bloomfield, New Jersey. Mr. Friend joined the Institute of Radio Engineers as an Associate in 1926 and transferred to Member Grade in 1938.

Clyde M. Hunt

The chairmanship of the Engineering Committee for the Fourth District of the National Association of Broadcasters has been accepted by Clyde M. Hunt, chief engineer for station WTOP, Columbia-owned and -operated station for Washington, D. C. In this capacity, he will co-ordinate the activities on behalf of the industry of chief engineers of member stations.

Mr. Hunt joined the Institute of Radio Engineers as an Associate in 1934 and transferred to Member Grade in 1941. At the present time he is chairman of the Washington section of the I.R.E.

I. J. Kaar

I. J. Kaar has been appointed manager of the receiver division, of General Electric's electronics department at the company's Bridgeport, Connecticut, works.

Mr. Kaar, who formerly was managing engineer of the receiver division, was graduated from the University of Utah in 1924 with a B.S. degree in electrical engineering. In October of that year he joined General Electric at Schenectady as a student engineer on "test," and in 1925 was transferred to the radio engineering department where he was engaged for several years on the development and design of high-power transmitters. In January, 1933, he entered the general engineering laboratory to work on the development of radio receivers, and made many contributions to vacuum-tube and radio-circuit design.

In September of 1934 Mr. Kaar was transferred with the nucleus of what is now the receiver division engineering section to the General Electric Bridgeport plant. On November 15 of the same year he was appointed designing engineer of the radio receiver section there. Later, when the radio and television (now electronics) department was formed, Mr. Kaar became designing engineer of the receiver division of that department. On October 1, 1941, he was named managing engineer of the receiver division, the position he held until his new appointment.

Mr. Kaar is a native of Dunsmuir, California. He is a member of the Society of Naval Engineers, Sigma Nu, and Tau Beta Pi, a Fellow and past director of the Institute of Radio Engineers, and chairman of the television technical committee of that body.

A. H. Rosenthal

A. H. Rosenthal, physicist and electronic engineer, has been appointed director of research and development of Scophony Corporation of America, according to an announcement made by Arthur Levey, president of Scophony, with which are associated Television Productions, Inc. (a Paramount subsidiary), and General Precision Equipment Corporation.

Dr. Rosenthal was connected with Scophony, Ltd., of London for several years, and in his present position will head a group of scientists and engineers engaged in research and development of inventions, not only in television, but also in the field of electronics, including various applications of supersonics.

Russell H. Varian

In recognition of his contribution to the development of radio location, and particularly with reference to the invention and evolution of the klystron, the Brooklyn Polytechnic Institute of Brooklyn, N. Y., in June, 1943, conferred the honorary degree of Doctor of Science on Russell H. Varian, research associate at Stanford University. Mr. Varian is an Associate Member of The Institute of Radio Engineers.

Ray Zender

In addition to his duties as Chief Engineer and Sales Manager of Lenz Electric Manufacturing Company, Chicago, wire manufacturers, Ray Zender, an Associate Member of the Institute, has been appointed wire consultant to the Radio and Radar section of the War Production Board on a "dollar-a-year" basis.

Correspondence

The following method is suggested for the solution of transient circuits whose steady state can be written by J. Millman's network theorem.[1] The problem is to determine the voltage across two points when a constant voltage is applied across two other points. This response voltage can be expressed as an admittance function, solvable by operational methods.

Millman shows that the steady-state voltage V_{00}, of a system of n admittance branches Y_K, terminating at 0, where the voltage V_{0K} at each of the other ends (K) of the admittances is known, is

$$V_{00'} = \frac{\sum_{K=1}^{M} V_{0K} Y_K}{\sum_{K=1}^{M} Y_K}, K = 1, 2, 3, \cdots M. \quad (1)$$

[1] J. Millman, "A useful network theorem", Proc. I.R.E., vol. 28, p. 413; September, 1940.

Reducing (1) for a single constant voltage V_{01} for the transient solution,

$$V_{00'} = \frac{V_{01} Y_1}{\sum_{K=1}^{M} Y_K}. \quad (2)$$

This is the form of the response *current* as systematized by Heaviside. The solution can be found by using a Heaviside formula or the Laplacian method. A simple example will illustrate this process.

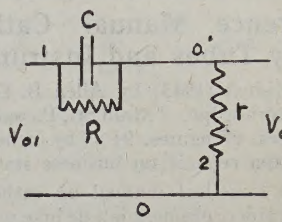


Fig. 1

In Fig. 1 the voltage $V_{01}=0$ when $t<0$, and constant when $t>0$ produces the response voltage $V_{00'}$:

$$V_{00'} = \frac{V_{01}(1/R + PC)}{1/r + 1/R + PC},$$

where PC is the operational admittance for C. By Heaviside's expansion theorem

$$\frac{g(P)}{h(P)} 1 = \frac{g(0)}{h(0)} 1 + \sum_{s=1}^{M} \frac{g(P_s) \exp(-P_s t)}{P_s g'(P_s)} 1.$$

Therefore,

$$V_{00'} = V_{01}\left[\frac{1/r}{1/r+1/R} + \frac{1/R-(1/r+1/R)}{-(1/r+1/R)} \exp\left(-\frac{t}{RC} - \frac{t}{rC}\right)\right]$$

or,

$$= V_{01}\left[\frac{r}{r+R} + \frac{R}{r+R} \exp\left(-\frac{t}{RC} - \frac{t}{rC}\right)\right].$$

Norman E. Polster
Research Department
Leeds and Northrup Co.
Philadelphia, Pa.

Quarterly of Applied Mathematics

A new periodical in the field of applied mathematics has appeared through the issuance, in April, 1943, of the first number of the *Quarterly of Applied Mathematics*. Its purpose is stated to be an attempt "to meet the needs of certain mathematicians and engineers whose interests extend beyond the accepted boundaries of their respective groups. These mathematicians find their greatest interest in the applications of mathematics to physical problems, and these engineers seek solutions of practical problems by advanced mathematical methods." As examples of subjects lying within the circle of interest of the new journal are mathematical theories related to engineering problems, e.g., in "fluid mechanics, elasticity, plasticity, thermodynamics," and certain phases of classical mechanics and electrical engineering. Of possible interest to radio-and-electronic engineers are papers in the first two issues of the *Quarterly* dealing with vibrations of a clamped plate under tension, the theory of direct image errors, the impedance of a transverse wire in a rectangular wave

guide, and forced vibrations of systems with nonlinear restoring force. The annual subscription price is $6.00, and orders may be addressed to the *Quarterly of Applied Mathematics*, Brown University, Providence 12, Rhode Island.

Books

Reference Manual—Cathode-Ray Tubes and Instruments

Published (1943) by Allen B. DuMont Laboratories, Inc., 2 Main St., Passaic, N. J. 116 pages. 87 figures. $9\frac{1}{4}\times 11\frac{1}{2}$ inches. Free on written request on business stationery.

This loose-leaf manual on cathode-ray oscillograph equipment is a de luxe presentation of the engineering requirements of such equipment, especially when used in the highly specialized measurement and research fields. It particularly directs the attention of oscillograph users to the most effective utilization of the many features provided in a modern oscillograph.

It is combined with the technical data about the most popular items of DuMont equipment, embracing complete oscillographs, cathode-ray tubes, and certain associated items that are necessary in particular tests.

It gives the reader a complete review of what he needs to know in order to use an oscillograph as an item in his tests, and shows what models are most suitable for the job. The binding permits the inclusion of supplements when issued to cover future developments. This service will also, in all likelihood, include additions to the section "Application Notes," which bids to be a valuable feature of the manual.

Ralph R. Batcher
Hollis, L. I., N. Y.

Practical Radio for War Training, by M. N. Beitman

Published (1943) by Supreme Publications, 328 S. Jefferson St., Chicago, Ill. 332 pages+4-page index. 288 figures. 6×9 inches. Price, $2.95.

This book was written to present to those without previous radio training or knowledge the elementary theory of electricity and radio with special emphasis on radio repair, adjustment, and operation. It is entirely nonmathematical. Its appeal is primarily to the beginner, either in home-study or war-training courses. For this class of students it should prove useful. It is probable that those pursuing somewhat more advanced radio and electrical work who have not had practical experience will be able to profit by a reading of this book because of the emphasis which is placed on practical matters.

A fair idea of the nature and scope of the material covered can best be obtained by a consideration of the type of topic listed in the table of contents. The following list is not complete but is probably representative:

What Makes Up a Radio Receiver
Mechanics of Radio
Circuits Using Resistors
Properties of Coils and Transformers
Circuits
How Meters Work
Vacuum Tubes
Power Supplies
Audio Amplifiers and Accessories
Test Equipment Using Meters
Electronic Test Equipment, etc.

The book is well and generously illustrated and appears to be thoroughly up to date. The subjects of frequency modulation and television are not included. However, it is quite likely that a discussion of such subjects is beyond the scope of an elementary text.

W. O. Swinyard
Hazeltine Electronics Corporation
Chicago, Ill.

Basic Electricity for Communication, by W. H. Timbie

Published (1943) by John Wiley and Sons, Inc., 601 W. 26 St., New York 1, N. Y. 590 pages+13-page index+ix pages. 464 figures. $5\frac{3}{4}\times 8\frac{3}{4}$ inches. Price, $3.50.

The author has had years of experience as a teacher and as a writer of books on elementary electric-circuit theory as a result of which he is unusually well qualified for the task in hand. In introducing electrical principles he uses familiar illustrations from everyday experiences and presents the results in clear-cut, easily understandable English.

"Basic Electricity" will appeal to those having a limited mathematical foundation and especially to those who are trying to improve their usefulness through individual study. The treatment is adequate for most purposes and the many problems that are worked out in detail will prove helpful in familiarizing the reader with methods of solving problems. About half of the book is devoted to direct-current circuits, including ammeters, voltmeters, and Wheatstone bridges, and the remainder is divided about equally between alternating-current circuits and electronic phenomena including various applications of electronic tubes to communication purposes.

H. M. Turner
Yale University
New Haven, Conn.

High Frequency Thermionic Tubes, by A. F. Harvey

Published by John Wiley and Sons, Inc., 440 Fourth Avenue, New York, N. Y. 229 pages+6-page index+viii pages. 99 figures. $5\frac{3}{4}\times 8\frac{3}{4}$ inches. Price, $3.00.

The title of this book describes very well its scope and contents. It is hardly a textbook but will form a valuable addition to the library of every tube engineer working on short-wave tubes (and that includes nearly all of them these days).

There are only six chapters, the fundamentals being cleared away in the first chapter so that the body of the text deals with the subject in hand. Material on the magnetron covers nearly half the total pages of text and this proportion seems justifiable. Retarding field tubes and the Klystron and allied tubes are also adequately covered for the scope of the book.

In general, the text is of two sorts: (1) a brief résumé of the published material of workers in the field; and (2) details of the work of the author, largely of an experimental nature on certain aspects of his subject, such as use of positive ions for the study of retarding field tube phenomena, and several of the characteristics of the magnetron.

An excellent feature of the book is the bibliography. There are over 500 references and they are grouped together at the end of each chapter, which in some ways is more useful than cluttering up each page with them or putting them in one large list at the end of the book.

It is interesting, but regrettable, to note that of the 160 references included relating to the oscillating type of magnetron, the only ones listed prior to 1932 are by German or Japanese workers.

The author appears to have omitted the first disclosure, published in English, on the split-anode oscillating magnetron. This appeared in the June, 1928, Proceedings of the I.R.E. under the title "Beam Transmission of Ultra Short Waves." The author was H. Yagi. Part II of this paper is devoted to oscillating magnetrons capable of producing oscillations as low as 12 centimeters in wavelength. In a number of respects, this was a prophetic paper.

American readers may not like the unfamiliar British-tube-type numbers particularly under pictures, some of which are of well-known American types.

The author groups oscillating magnetrons into three classes or "regimes" as he names them. One of these, he terms the "dynatron regime." This may be confusing to American readers who have come to associate this word with the utilization of secondary emission. The author, however, uses it in connection with a negative resistance characteristic rather than secondary emission.

One cannot help but admire the courage of any author who will publish in the year 1943 a book on "high-frequency tubes." He must either have been well insulated from recent military applications in this field or, as is more likely, been severely limited in what he could present to his readers from the material at hand.

This book will probably be as much up to date on its subject as is possible for the duration and as such will be found useful by radio and also, as we now can say, radar engineers.

W. C. White
General Electric Company
Schenectady, N. Y.

Radio Troubleshooter's Handbook (Third Revised Edition), by Alfred A. Ghirardi

Published by Radio and Technical Publishing Co., 45 Astor Place, New York 3, N. Y. 738 pages+6-page index+viii pages. 112 figures. $8\frac{1}{2}\times 11\frac{1}{2}$ inches. Price, $5.00.

This is a working manual prepared from radio receiver case histories, for the assistance of radio repairmen in trouble shooting, adjusting, and making repairs. This third edition has been greatly enlarged in the

amount of material compiled and contains 190 pages of reference data on tubes and other radio components of value to repairmen, in addition to the specific data on receiver set troubles.

A radio engineer may have a hard time justifying the method of repairing receivers on which this book is based in that it replaces the importance of knowledge of principles with experience—more specifically the experience of others repairing those same receiver models. It might be a good thing if this handbook became required reading for those design engineers and others involved in the production of sets that have the worst case histories, for pointing out how not to do things, in order to improve a receiver's reliability in the future.

RALPH R. BATCHER
Hollis, L. I., N.Y.

Radio Engineers' Handbook, by Frederick Emmons Terman

Published (1943) by McGraw-Hill Book Company, 330 W. 42 St., New York 18, N. Y. 995 pages+23-page index+xi pages. 869 figures. 6×9 inches. Price, $6.00. Bound in maroon imitation leather, semiflexible covers.

Prepared by one of the most eminent authors of radio engineering textbooks, this reference volume is distinguished by the same thorough research and masterful presentation as his previous works. Compiling a handbook is a difficult problem of space allotment and the selection of material; the author has done well in concentrating on the aim of completeness with brevity, so the result is not only the most up-to-date but also the best organized handbook in the radio field.

The presentation is directed to radio engineers who have had college training or special radio courses. While little space is devoted to theory and textbook explanations, the basic principles are summarized as an introduction to the information presented by formulas, curves, charts, and tables. This gives the reader much more confidence in using the material, and the extensive citations to other publications give him ample guidance for verifying the information in the handbook or studying further into the subject. The magnitude of the work is indicated by the fact that it contains material selected from 1500 references cited and other sources too numerous to list. Every section is not only a short course of study but an inspiration to further investigation. Departing from the usual practice of having each section written by a different specialist, the preparation of this handbook by a single author has done much to unify the style and to minimize duplication of subject matter. The difficulty of different systems of units still remains, but no simple solution of this problem has been found.

The scope is similar to that of earlier radio handbooks. After a brief mathematical introduction as an aid in using the formulas, there is much information on the design of circuit components of all kinds.

The section on circuit theory is an excellent compilation of the principal theorems and design formulas of resonant circuits and wave filters, even to the basic relations between amplitude and phase.[1] This is accompanied by a very good treatment of transmission lines and one of the first concise summaries of wave guides and cavity resonators. The latter includes tables and curves of the cutoff frequency and attenuation in wave guides, and the Q and resonant impedance of cavity resonators. The companion subject of horns is well treated in another section with antennas

Vacuum tubes and their direct applications occupy more than one third of the volume. The treatment of the various types of tubes includes an unusual collection of design information on electron lenses. In the section on amplifiers, most attention is paid to audio-frequency amplifiers, their amplification and phase characteristics. Video-frequency and radio-frequency amplifiers also are well treated. All kinds of oscillators, especially ultra-high-frequency oscillators, are covered in another section, followed by modulators, detectors, and power rectifiers.

One fifth of the book is devoted to wave propagation and antennas. In many ways, this is the outstanding contribution of the Handbook, because such a careful selection and correlation has been required to give the reader a fairly complete collection of data while minimizing duplication and confusion among the various original authors. The data on wave propagation over the earth, which have been published by K. A. Norton and C. R. Burrows, are well summarized. This is followed by a treatment of the ionosphere and of the factors governing the choice of an operating frequency. In the field of antennas, special attention is paid to radiation resistance and mutual impedance of various systems, and to the directivity of arrays, horns, and reflectors. Many types of antennas are described, concluding with the wide-band antennas used for television.

As the author explains, the intended section on television would have delayed the book beyond the period of war work, so the publication was not held up for the completion of this material.

After a brief introduction to navigation aids, the Handbook closes with a brief outline of the more common measurements on circuits and tubes.

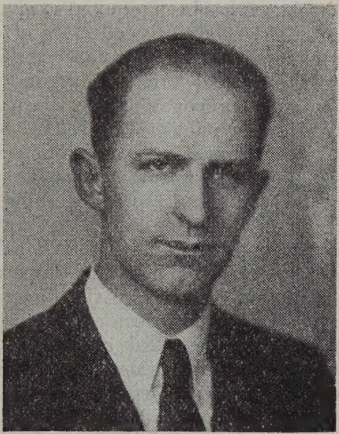
The practical viewpoint of the author is exemplified by his having given a favored location (inside the back cover) to the reactance chart, which has become the most essential single piece of reference material used by both practical designers and research workers. On the other hand, no attempt has been made to supplant the excellent handbooks of vacuum-tube manufacturers in this field where new types and new applications are developing so rapidly.

This Handbook is recommended as a welcome addition to the library of every radio engineer who is alert to the advances in his profession and who is anxious to accomplish the best work within limited time.

H. A. WHEELER
Hazeltine Electronics Corporation
Little Neck, L. I., N. Y.

[1] F. E. Terman, "Network theory, filters, and equalizers", PROC. I.R.E., vol. 31, pp. 164-175, 233-240; and 288-304; April, May and June, 1943. (This is a preprint of a section from the Handbook.)

Contributors



W. F. Goetter

W. F. Goetter (A'41) was born at Hillsboro, Oregon, in 1912. He joined the General Electric Company shortly after receiving the B.S. degree in electrical engineering from Oregon State College in 1937. Upon completion of the test course, Mr. Goetter entered the radio transmitter engineering department and became intimately associated with the development of frequency-modulation broadcast equipment. Since the beginning of the war his work has been on secret apparatus for the armed forces. He is a member of Eta Kappa Nu, Phi Kappa Phi, Tau Beta Pi, and an associate member of Sigma Xi.

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Raymond A. Heising (A'20-F'23) was born on August 10, 1888, at Albert Lea, Minnesota. He received the E.E. degree in 1912 from the University of North Dakota and the M.S. degree in 1914 from the University of Wisconsin. Since 1914 he has been a member of the technical staff of the engineering department of the Western Electric Company and its successor, the Bell Telephone Laboratories. Mr. Heising developed and constructed the experimental transoceanic radiotelephone transmitter used at Arlington in 1915 and the constant-current modulation system used almost exclusively in World War I radiotelephones and in early broadcast transmitters. During World War I, he developed radiotelephone sets for the United States Army and Navy and educated in his laboratory men who became radio communication instructors in the Signal Corps. Since the war, he has continued research and development work in

Raymond A. Heising

connection with ship-to-shore operation, short- and long-wave transoceanic circuits, ultra-short waves, piezoelectric devices, War II problems, etc. He was awarded the Morris Liebmann Memorial Prize in 1921 and was President of the Institute in 1939.

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Stanford C. Hooper (F'28-A'33) was born on August 16, 1884, at Colton, California. He was graduated from the United States Naval Academy in 1905; became an Ensign in 1907, Lieutenant in 1910, Lieutenant Commander in 1916. Commander in 1921, Captain in 1927, and Rear Admiral in 1938. During World War I, he served on various naval vessels. He instructed in electricity, physics, and chemistry at the Naval Academy from 1910 to 1911. From 1912 he served for two years as the first Fleet Radio

Stanford C. Hooper

Officer, resuming that post again from 1923 to 1925. For eleven years between 1914 and 1928 he was in charge of the Radio Division of the Navy Department and later served as Director of Naval Communications until 1935 when he was transferred as Director of the Technical Division, under the Chief of Naval Operations. In 1942 he was appointed General Consultant for Radio and Underwater Sound for the Navy.

Admiral Hooper was the first technical adviser to the Federal Radio Commission and he attended various national and international radio conferences. He was awarded the Institute Medal of Honor in 1934 for the orderly planning and systematic organization of radio communication in the Government Service with which he was associated, and the concomitant and resulting advances in the development of radio equipment and procedure. He has also been awarded the following decorations and medals: Navy Cross; Mexican Service Medal; Victory Medal, Destroyer Clasp; American Defense Service Medal; and the Legion of Honor.

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For a biographical sketch of Ronold King see the Proceedings for October, 1943; for Robert I. Sarbacher, August, 1943.